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SECTION 22

POWER ELECTRONICS

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22.1 INTRODUCTION

22.1.1 Role of Power Electronic Converters

Power electronics is an enabling technology that achieves conversion of electric power from one form to another, using a combination of high-power semiconductor devices and passive components—chiefly transformers, inductors, and capacitors. The input and output may be alternating current (ac) or direct current (dc) and may differ in magnitude and frequency. The conversion sometimes involves multiple stages with two or more converters connected in a cascade. The end goals of a power electronic converter are to achieve high efficiency of conversion, minimize size and weight, and achieve desired regulation of the output. Figure 22-1 shows power electronic converters in a generic application.

22.1.2 Application Examples

Power electronic converters can be classified into four different types on the basis of input and output, *dc-dc*, *dc-ac*, *ac-dc*, and *ac-ac*, named with the first part referring to the input and the second to the output. The diode bridge rectifier is the front end for most low-power converters. It converts line frequency ac (e.g., from a wall outlet) to an unregulated dc voltage, and the process is commonly called rectification. In a *dc-dc* converter, both the input and the output are dc, and in the simplest case the output voltage needs to be regulated in presence of variation in load current and changes in the input voltage. A computer power supply has a diode bridge front end followed by a dc-dc converter, the combination of which converts line frequency ac voltage to several regulated dc voltages (Fig. 22-2). Electronic ballasts for compact fluorescent lamps consist of a line frequency rectifier followed by a dc to high-frequency ac converter (frequency range of 20 to 100 kHz) whose output is connected to a resonant tank circuit that includes the load. In an adjustable speed motor drive application (Fig. 22-3), the input is a 3-phase ac supply, and the output is a 3-phase ac whose magnitude and frequency are varied for optimum steady-state operation and dynamic requirements of the drive.

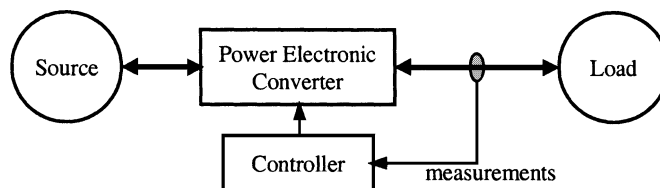


FIGURE 22-1 Application of power electronic converters.

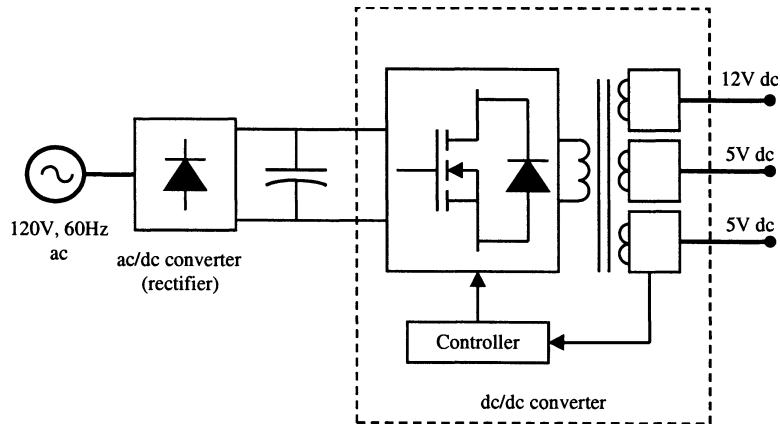


FIGURE 22-2 Computer power supply.

Development of power semiconductors with very high voltage and current ratings has enabled the use of power electronic converters for utility applications. In transmission systems, power electronic converters are being utilized to control power flow, damp power oscillations, and enhance system stability. At the distribution level, power electronic converters are used for enhancing power quality by means of dynamic voltage restorers, static var compensators, and active filters. Power electronic converters also play a significant role in grid connection of distributed generation and especially renewable energy sources; their functions include compensation for steady state and dynamic source characteristics leading to optimal energy transfer from the source, and protective action during contingencies.

Future automobiles are expected to have a large number of power electronic converters performing various functions, for example, electric power steering, active suspension, control over various loads, and transferring power between the conventional 14-V bus and the recently proposed 42-V *Power Net* [1]. Hybrid electric and all-electric vehicles also utilize controlled power electronic converters for interfacing the battery and motor/generator.

The proliferation of power electronics connected to the utility grid has also led to power quality concerns due to injection of harmonic currents by grid-connected inverters, and highly distorted currents drawn by diode bridge rectifiers. Due to fast transients of voltages and currents within power

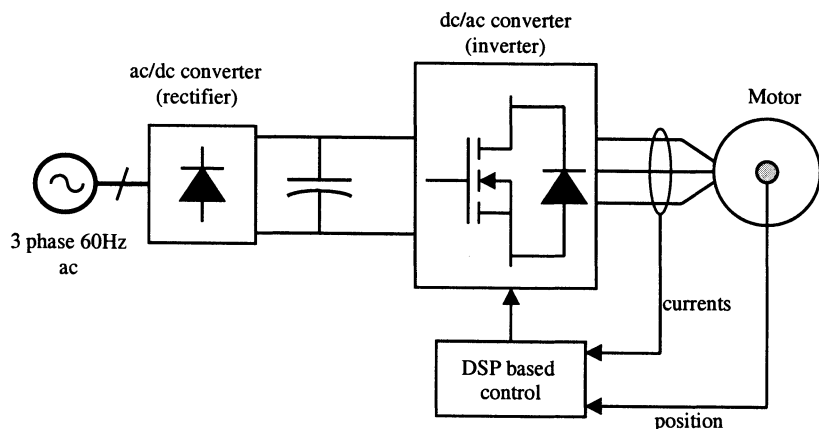


FIGURE 22-3 Adjustable speed motor drive.

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converters, they can be a source of electromagnetic emissions leading to electromagnetic interference. Several solutions to limit and correct these effects have therefore been developed.

22.1.3 Scope and Organization

This section gives an overview of power electronic systems. Details of specific converter types and applications have been omitted and only the fundamentals are presented. In some cases, important results are stated without derivation. Mathematical content has been kept to a minimum. In places, empirical aspects have been included, since power electronics is an application-oriented discipline. Design procedures are presented with only those justifications that were deemed imperative. A long list of references consisting of textbooks on the subject of power electronics, reference books on specific areas and applications of power electronics, important research publications, and several online sources has been provided. The reader is expected to use this section as a starting point, followed by the references on the topic of particular interest.

First, the basic principles for analysis and design of power converters are presented in Sec. 22.2. Topology and operating principles of the four types of power electronics converters are described with one section devoted to each. A very simple description of power electronic converter control is presented using the example of dc-dc converters. This is followed by deleterious effects of power electronic converters and precautions necessary to limit or correct them. Applications are described next bringing together the requirements and complete power electronic system realization for some specific examples. Finally, the individual components that constitute a power electronic converter are discussed. Current research initiatives and expected future trends are indicated in each section.

22.2 PRINCIPLES OF SWITCHED MODE POWER CONVERSION

This section presents some basic principles that are common to the analysis of all switch mode power converters. Line-commutated power electronic converters are not, strictly speaking, switched mode converters; they are discussed in Sec. 22.6.

22.2.1 Bipositional Switch

The most basic component of a switch mode power converter is the bipositional switch shown in Fig. 22-4a. Nodes 1 and 2 of the switch are invariably connected across a dc voltage source (or across

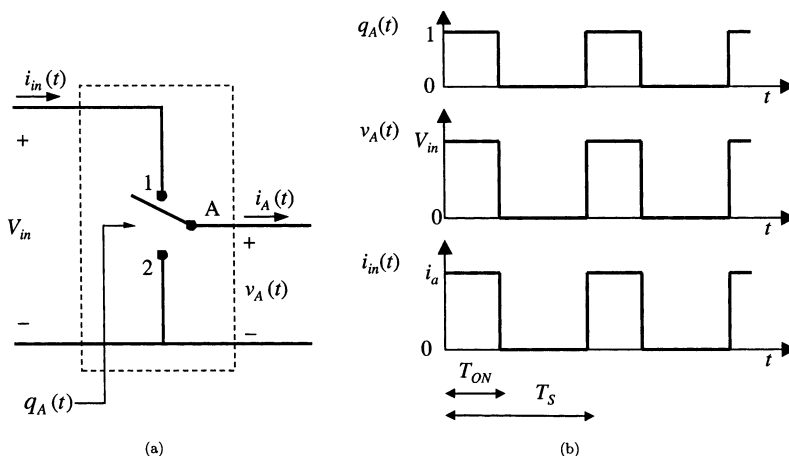


FIGURE 22-4 (a) Bipositional switch (b) switching waveforms.

TABLE 22-1 States of a Bipositional Switch

$q_A(t)$	Switch position	MOSFET & diode state	Pole voltage & input current
1	1	S1 & D1 ON, S2 & D2 OFF	$v_A = V_{in}$, $i_{in} = i_A$
0	2	S1 & D1 OFF, S2 & D2 ON	$v_A = 0$, $i_{in} = 0$

a big capacitor whose voltage is close to a constant dc), and pole “A” of the switch is in series with a dc current source (or a big inductor whose current is close to a constant dc). This bipositional switch, which is also referred to as a switching power pole, switches at very high frequencies, and is controlled by the signal $q_A(t)$. The switched pole A voltage and the input current based on the control signal $q_A(t)$ are listed in Table 22-1, and the corresponding waveforms are shown in Fig. 22-4b.

Figure 22-5a shows the electronic implementation of a complete bipositional switch using metal-oxide-semiconductor field-effect transistors (MOSFETs). This implementation can support pole current in either direction and is useful for applications where current direction can reverse. In most dc-dc converter applications, the current through the pole A is unidirectional, and hence, the implementation shown in Fig. 22-5b is sufficient to realize the bipositional switch.

22.2.2 Pulse Width Modulation

The concept of pulse width modulation (PWM) is central to all switch mode power converters. Pulse width modulation refers to the control of the average value of a switching variable, for example, $v_A(t)$ in Fig. 22-4b, by controlling or modulating its pulse width. Some basic concepts and definitions necessary to understanding PWM are presented here.

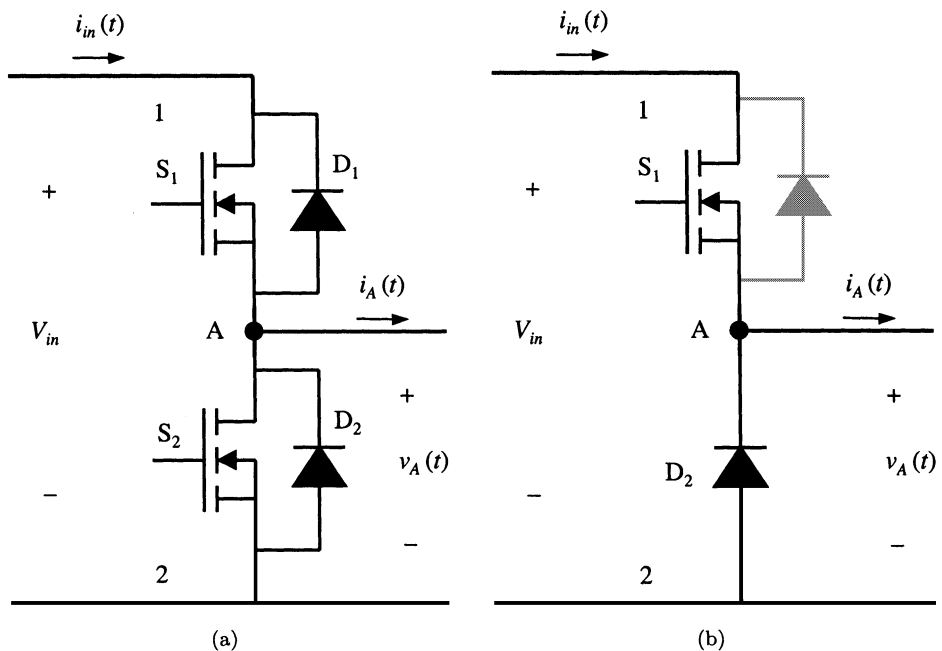


FIGURE 22-5 Electronic implementation of bipositional switch: (a) for bidirectional pole current (b) for unidirectional pole current.

Duty Ratio. The frequency at which the bipositional switch is switched on and off is denoted by f_s , and the corresponding time period by $T_s (= 1/f_s)$. The transition between the two states of the switch occurs in a very small duration compared to T_s . The time for which the switch remains in position 1 during a switching period is denoted by T_{on} . The duty ratio d of the bipositional switch is then defined as the ratio of on-time to total time period:

$$d = \frac{T_{on}}{T_s} \quad (22-1)$$

Averaging. Currents and voltages in power electronic converters have (1) high-frequency components corresponding to the switching frequency of the bipositional switch elements, and (2) low-frequency components due to slower variations caused by change in load demand, source magnitude, and changes in reference value of the desired outputs. For dynamic control and steady-state analysis, the low-frequency components are of primary interest. To study these components, it is sufficient to study their averages over one switching time period. It should be noted that the averaging presented here [2] is a very basic form of the general averaging method [3] and has limitations in terms of validity with respect to the switching frequency. However, this simplification is good enough for most practical purposes, and can be confidently used for steady state and dynamics up to one-fifth the switching frequency. Throughout this chapter the averaged variables, that is, averaged over one switching period, are denoted by a “-” on top of the variables. Thus, the averaged value of $x(t)$ is given by

$$\bar{x}(t) = \frac{1}{T_s} \int_{t-T_s}^t x(\tau) d\tau \quad (22-2)$$

In steady state, the average values of $q_A(t)$ and $v_A(t)$ are given by

$$\bar{q}_A = \frac{1}{T_s} \int_0^{T_s} q_A(\tau) d\tau = \frac{1}{T_s} \int_0^{T_{on}} 1 d\tau = \frac{T_{on}}{T_s} = d \quad (22-3)$$

$$\bar{v}_A = \frac{1}{T_s} \int_0^{T_s} v_A(\tau) d\tau = \frac{1}{T_s} \int_0^{T_{on}} V_{in} d\tau = d \cdot V_{in} \quad (22-4)$$

In general, the averaged quantities can be time varying, since the pulse widths of the switching waveform can vary with time. Thus

$$\bar{q}_A(t) = d(t) \quad (22-5)$$

$$\bar{v}_A(t) = d(t) \cdot V_{in} \quad (22-6)$$

As an example of PWM, we can regulate the average value of $v_A(t)$ in Fig. 22-4b by varying the duty ratio d . If $V_{in} = 10 \text{ V}$, $f_s = 100 \text{ kHz} \Rightarrow T_s = 10 \mu\text{s}$, then $T_{on} = 5 \mu\text{s} \Rightarrow d = 0.5$, and $v_A = 5 \text{ V}$, etc. By varying the duty ratio sinusoidally a low-frequency ac voltage can be synthesized from a dc voltage, as illustrated in Fig. 22-6.

22.2.3 Concept of Steady State

A converter is said to be in dc steady state when all its waveforms exactly repeat in each switching period, that is, $x(t) = x(t - T_s) \forall t$, where x is any of the converter variables. With reference to Eq. (22-2), it is clear that in steady state the average value of any variable is constant. Analysis of steady-state operation is essential to determine ratings and design of the power stage components in the converter, viz, power semiconductor devices, inductor, capacitors, and transformers. Important

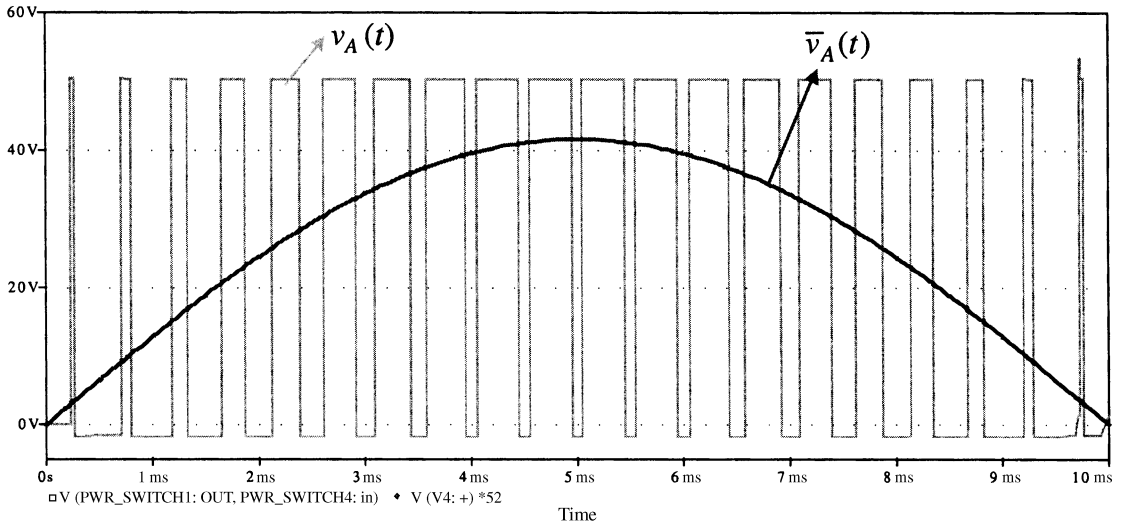


FIGURE 22-6 AC synthesis using PWM.

concepts that enable steady-state analysis from a circuit view point are discussed below. It should be remembered that these are only valid during steady-state operation.

Steady-State Averages of Inductor Voltage and Capacitor Current. The instantaneous v - i relationship for an inductor is

$$v_L(t) = L \frac{di_L(t)}{dt} \quad \text{or} \quad i_L(t) = i_L(0) + \frac{1}{L} \int_0^t v_L(\tau) d\tau \quad (22-7)$$

where $v_L(t)$ is the voltage across the inductor and $i_L(t)$ is the current flowing through the inductor. Since $i_L(T_s) = i_L(0)$ in steady state, from the integral form of Eq. (22-7) it follows that

$$\bar{v}_L = \frac{1}{T_s} \int_0^{T_s} v_L(\tau) d\tau = 0 \quad (22-8)$$

The above relationship can also be derived directly in terms of the averaged quantities as follows:

$$\bar{v}_L(t) = L \frac{d\bar{i}_L(t)}{dt} = 0 \quad (\text{since } \bar{i}_L(t) \text{ is constant in steady state}) \quad (22-9)$$

This is referred to as *volt-second balance* in an inductor. Figure 22-7 shows a typical steady-state waveform of an inductor voltage for many power converters. The positive area is exactly cancelled by the negative area, making the average value zero. It may be mentioned that during the start-up transient, $\bar{v}_L$ remains positive for several switching cycles, allowing the inductor current to rise from zero to its final steady-state value.

In a similar fashion, it can be shown that in steady state the average current through a capacitor

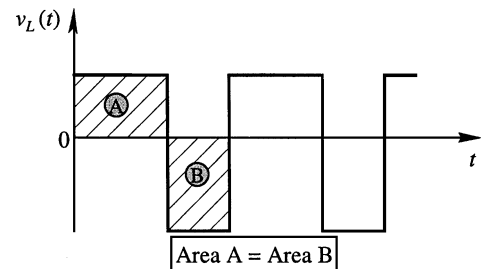


FIGURE 22-7 Volt-second balance for an inductor.

is zero. This is referred to as *ampere-second balance* in a capacitor. Note that though the average value of the capacitor current is zero, its root mean square (RMS) value, which is one of the main selection criteria for a capacitor, can be substantial depending on the converter topology.

Power Balance. For analytical purposes, it is often useful to neglect all losses in the converter and consider input power to be equal to the output power, again in an average sense

$$P_{in} = \bar{v}_{in} \bar{i}_{in} = P_o = \bar{v}_o \bar{i}_o \quad (22-10)$$

This implies that there is no increase or decrease in the energy stored in inductors and capacitors over one switching time period. This is valid for the input-output of the entire converter as well as any intermediate stage.

Kirchoff's Laws for Averages. Just like the instantaneous quantities, the averaged quantities also obey Kirchhoff's current and voltage laws. The sum of average currents entering a node is zero. The proof follows from interchanging the order of summation (for individual currents) and integration (over a switching time period)

$$\sum_k \bar{i}_k = \frac{1}{T_s} \sum_k \left[\int_0^{T_s} i_k \right] = \frac{1}{T_s} \int_0^{T_s} \left[\sum_k i_k \right] = 0 \quad (\text{since } \sum_k i_k \equiv 0) \quad (22-11)$$

Similarly, the sum of average voltages in a circuit loop is zero.

$$\sum_k \bar{v}_k = 0 \quad (22-12)$$

22.2.4 Power Loss in the Bipositional Switch

Electronic implementations of the bipositional switch shown in Figs. 22-5a and 22-5b have significant power loss. The power loss can be divided into two kinds—conduction loss and switching loss.

With reference to Fig. 22-5b, when the MOSFET is on there is a nonzero voltage across it. Similarly the diode has a forward voltage drop while it is conducting. Both of these lead to power loss whose sum averaged over one switching time period is called conduction loss.

A finite time interval is required to transition from one state to the other: (MOSFET on and diode off) to (MOSFET off and diode on), and vice versa. While the MOSFET is turning off, the diode cannot conduct until it is forward biased. As the voltage across the MOSFET increases from near zero to the full input voltage V_{in} , it conducts the full output current. Once the diode is forward biased the current starts transferring from the MOSFET to the diode. During the reverse transition, first current is transferred from the diode to the MOSFET, and then the voltage across the MOSFET reduces from V_{in} to the conduction voltage drop. Thus, the MOSFET incurs significant power loss during both transitions. The above description is simplified and there are other phenomena which contribute to loss during the transitions. The diode also has power loss during the transitions. The sum of losses in the MOSFET and diode during the transitions averaged over one switching time period is called the switching loss. Switching power loss increases with increase in switching frequency and increase in transition times. Sum of the conduction and switching loss, computed as averages over one complete switching periods, gives the total power loss Fig. 22-8.

Similar losses occur in the realization of Fig. 22-5a. When S_1 is turned off by its control signal, current $i_A(t)$ transfers to D_2 , the antiparallel diode of S_2 . After this transition, S_2 is turned on and the current transfers from the diode to the MOSFET channel (which can conduct in either direction). A short time delay, called dead time, is required between the on signals for S_1 and S_2 . The dead time prevents potential shorting of the input voltage, also known as shoot-through fault.

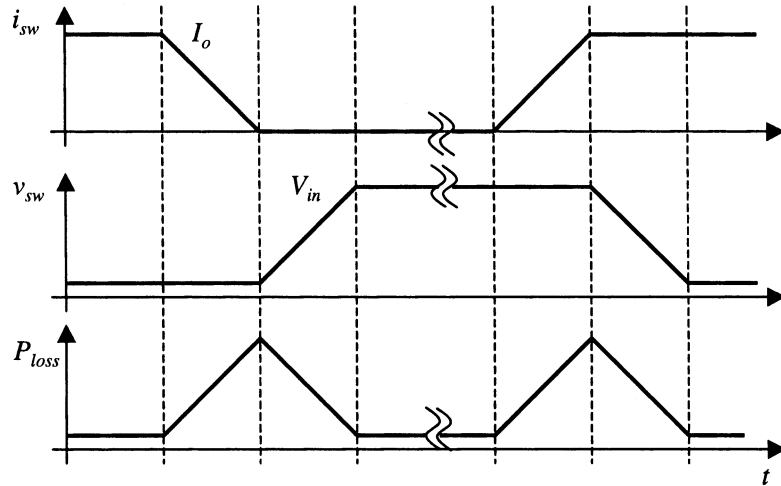


FIGURE 22-8 Switching transients in bipositional switch implementation.

The nonidealities of nonzero voltage drop and switching times will be neglected for analysis of power electronic converters presented throughout this chapter. However, these are extremely important in design and selection of components for a real power converter.

22.3 DC-DC CONVERTERS

DC-DC converters represent one major area in power electronics. In a dc-dc converter, the input and output may differ in magnitude, the output may be electrically isolated from the input, and the output voltage may have to be regulated in the presence of variation in input voltage and load current. In a typical power distribution system (for digital systems), several lower magnitude dc voltages are derived from a common input using a one or more converters. Battery-powered portable devices use converters that boost the input 1.5 V cell voltage to 5 or 9 V. Most of these converters have unidirectional power flow—from input to output. The presentation here is limited to the basic converter types. The interested reader is referred to text books that deal with details of these converters [4–8].

22.3.1 Buck Converter

The buck converter is used to step down an input voltage to a lower magnitude output voltage. Figure 22-9a shows the schematic of a buck converter. A power MOSFET and diode combination is shown for implementation of the bipositional switch with unidirectional output current. The bipositional switch is followed by an L-C low-pass filter that attenuates the high-frequency switching component of the pole A voltage and provides a filtered dc voltage at the output. A high switching frequency is desirable to reduce the size of the filter, the higher limit depending on the power level of the converter and the semiconductor devices used. The final choice of switching frequency depends on several factors: size, weight, efficiency, and cost. It is usually above the audible range and frequencies above 100 kHz are very common.

Operation. The input voltage V_{in} is assumed to remain constant within a switching cycle. The inductor L and capacitor C are sufficiently large so that the inductor current i_L and output voltage v_o do not change significantly within one switching cycle. The load is represented by the resistor R_L . Under

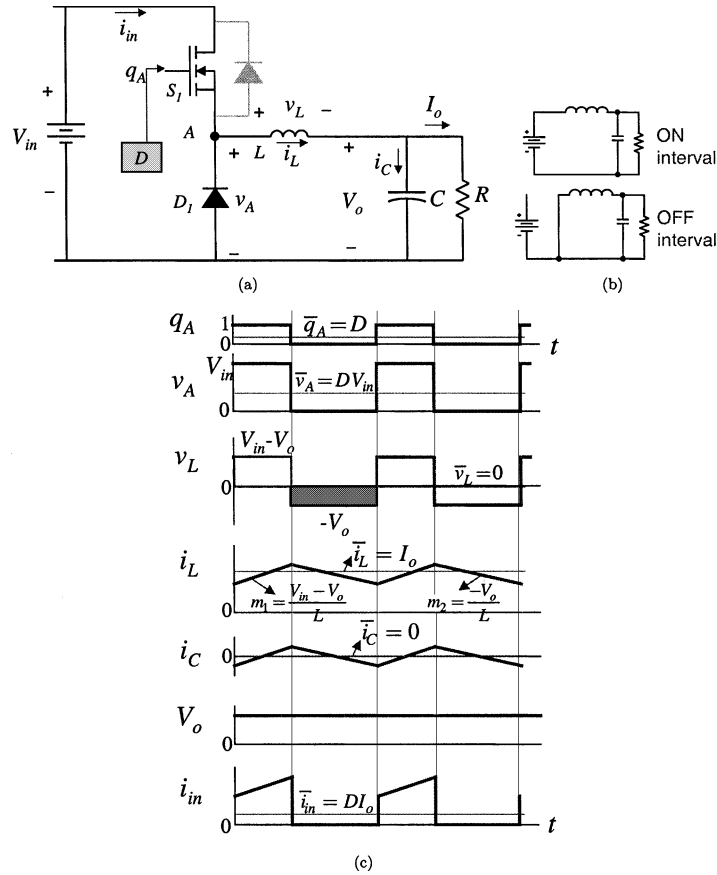


FIGURE 22-9 Buck converter: (a) circuit, (b) equivalent circuits during ON and OFF intervals, (c) steady-state waveforms.

steady-state operation it is assumed that the inductor current is always greater than zero. The MOSFET is turned on in response to signal $q_A(t)$ for $T_{on} = DT_s$, where D represents the steady-state duty ratio. During this time $v_A = V_{in}$ and $i_{in} = i_L$. When the MOSFET is turned off, the inductor current flows through diode D_1 leading to $v_A = 0$ and $i_{in} = 0$. Since the average voltage across the inductor is zero, $\bar{v}_L = 0$, the average output voltage is given by

$$\bar{v}_o = \bar{v}_A = DV_{in} \quad (22-13)$$

The average current through the capacitor C , $\bar{i}_C$, is zero. Thus, $\bar{i}_L = I_o$ and the input current is given by

$$\bar{i}_{in} = DI_o \quad (22-14)$$

From the above equations it is clear that the output voltage is lower than the input voltage and output current is higher than the input current. Also, power balance for averaged quantities can be verified from Eqs. (22-13) and (22-14). Within a switching cycle, instantaneous values of the inductor current and capacitor voltage vary as follows:

$$\begin{aligned} \text{MOSFET on:} \quad L \dot{i}_L &= V_{in} - v_o & C \dot{v}_o &= i_L - v_o/R_L \\ \text{MOSFET off:} \quad L \dot{i}_L &= -v_o & C \dot{v}_o &= i_L - v_o/R_L \end{aligned} \quad (22-15)$$

Equivalent circuits for the two intervals and instantaneous waveforms are shown in Figs. 22-9b and 22-9c.

Component Selection. Usually the inductor and capacitor are significantly large so that within a switching period v_o can be assumed constant in computation of i_L . This leads to the linear variation of i_L shown in Fig. 22-9c with a *peak-to-peak ripple* ΔI_L given by

$$\Delta I_L = \frac{V_o(1 - D)T_s}{L} \quad (22-16)$$

In most designs, the inductance value is chosen to limit ΔI_L between 10% and 30% of the full load current. Since the average capacitor current is zero, the instantaneous capacitor current is approximately equal to the ripple component of the inductor current.

$$i_C(t) = i_L(t) - I_o \quad (22-17)$$

The peak-to-peak capacitor voltage ripple resulting from the capacitor current can then be derived as

$$\Delta V_o = \frac{\Delta I_L \cdot T_s}{8C} \quad (22-18)$$

Capacitors used for filtering in most dc-dc converters are electrolytic capacitors, which are characterized by significant effective series resistance (ESR) and effective series inductance (ESL). These parasitics also contribute to the output voltage ripple and should supplement Eq. (22-18) in the choice of capacitors. Film or ceramic capacitors, which have significantly lower ESR and ESL, should be used in conjunction with electrolytic capacitors.

The MOSFET has to be rated to block a voltage greater than V_{in} , and conduct an average current greater than I_{in} . Power dissipation and temperature considerations usually require MOSFETs to be rated from 2 to 3 times the maximum input average current. In addition, the peak MOSFET current, equal to the maximum peak of the inductor current, should not exceed its maximum current rating. The diode has to be rated to block V_{in} , and conduct an average current greater than the maximum output current. Diodes are usually chosen with ratings approximately 2 times the expected maximum current.

PWM Control Implementation. As evident from Eq. (22-13) the duty ratio of the switch controls the output voltage. In response to variation in input voltage and load current, the duty ratio has to be changed by a feedback controlled system as shown in Fig. 22-10a. The error between the reference and actual output voltage is given to an appropriately designed error compensating amplifier, the output of which is a control voltage v_c . This control voltage is compared with a constant frequency sawtooth waveform. The output of the comparator is the switching signal $q_A(t)$ that determines the on or off state of the MOSFET. When the output voltage is lower than the reference value, the control voltage increases, leading to an increase in the duty ratio, which in turn increases the output voltage. The error amplifier and comparator, and several other features, are available in a single

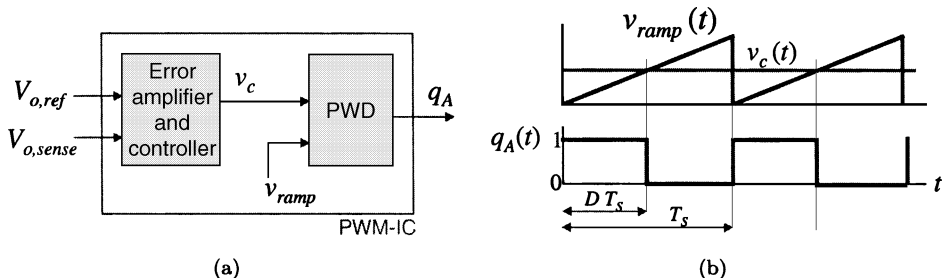


FIGURE 22-10 PWM generation: (a) ramp comparison, (b) control block diagram.

integrated circuit (e.g., UC3825A) available from several manufacturers (e.g., see Refs. [9–12]). This integration of components leads to reduction in overall size and cost.

22.3.2 Boost Converter

As evident from the name, the boost converter is used to step up an input voltage to a higher magnitude output voltage see Fig. 22-11a. In this case, the MOSFET is in the lower position while the diode is in the upper position. The inductor is on the input side and the output has a purely capacitive filter.

Assumptions made for analysis of buck converter are made here as well. When the MOSFET is on in response to $q_A(t) = 1$, diode D_1 is off, and the inductor current increases due to a positive voltage across it (Fig. 22-11b). When the MOSFET is switched off, the inductor current flows through diode D_1 and its magnitude decreases as energy is transferred from the inductor to the output capacitor and load. Instantaneous values of the variables during on and off intervals are

$$\begin{array}{llll} \text{MOSFET on} & v_A = 0 & i_d = 0 & Li_L = V_{in} \quad C\dot{v}_o = -v_o/R_L \\ \text{MOSFET off} & v_A = v_o & i_d = i_L & Li_L = V_{in} - v_o \quad C\dot{v}_o = i_L - v_o/R_L \end{array} \quad (22-19)$$

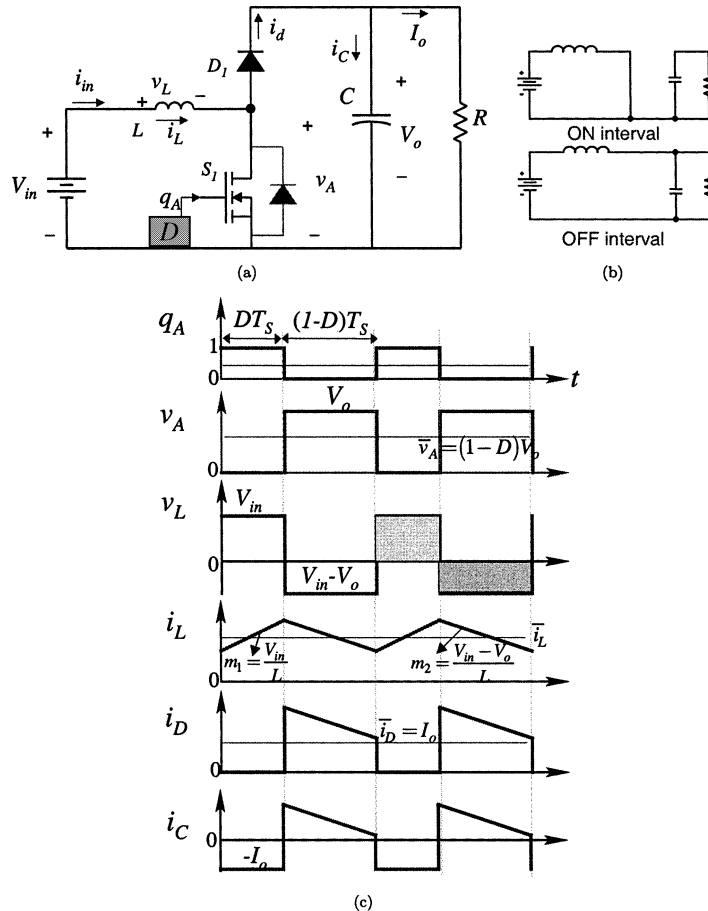


FIGURE 22-11 Boost converter: (a) circuit, (b) operating states, (c) steady-state waveforms.

Noting that $\bar{v}_L = 0$, the averaged pole A voltage is $\bar{v}_A = V_{in} = (1 - D)V_o$. In addition, using $\bar{i}_C = 0$, the steady-state conversion ratios for the boost converter can be obtained as follows:

$$\frac{V_o}{V_{in}} = \frac{1}{1 - D} \quad \text{and} \quad \frac{I_{in}}{I_o} = \frac{1}{1 - D} \quad (22-20)$$

From the above equation it is evident that the output voltage is always higher than the input voltage, and conversely, the output current is always lower than the input current by the same ratio. Waveforms of the boost converter variables are shown in Fig. 22-11c. The PWM implementation is the same as in the buck converter (Fig. 22-10), with the control objective being regulation of output voltage to a desired value.

The buck and boost converters are capable of either decreasing or increasing the input voltage magnitude, but not both. The buck-boost converter is the third basic converter that can be used to obtain an output voltage both lower and higher than the input voltage; since the output voltage is usually maintained constant, this implies that the input voltage may be higher or lower than the output voltage. A drawback of the buck-boost converter is that the output voltage polarity is inverted with respect to the input voltage return. The SEPIC converter (single-ended primary inductor converter) provides buck and boost gain without polarity inversion but at the expense of additional components. The Ćuk converter, derived from the buck-boost converter using the duality of current and voltage, is another basic dc-dc converter topology [4, 5]. None of the above converters have electrical isolation between the input and the output; however, isolated versions for all of these can be derived.

22.3.3 Flyback Converter

Figure 22-12a shows the buck-boost converter circuit. Discussion of this converter in its original configuration is omitted here. Instead, its electrically isolated version known as the flyback converter is described. The flyback converter is very common for low power applications. It has the advantage of providing electrical isolation with low component count. Derivation of the flyback converter from the buck-boost converter is shown in Fig. 22-12a. The flyback converter has a coupled inductor instead of an inductor with just one winding. The primary winding is connected to the input while

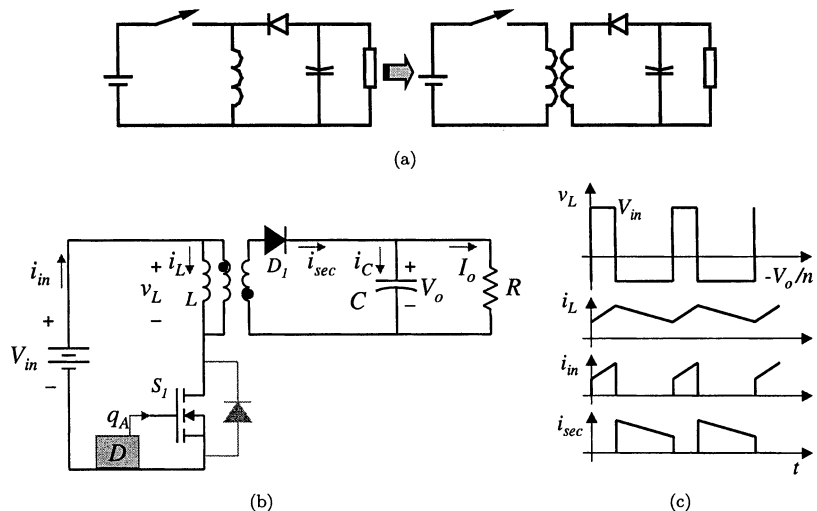


FIGURE 22-12 Flyback converter: (a) derivation from buck-boost, (b) circuit, (c) steady-state waveforms.

the secondary is connected to the output. The circuit diagram is shown in Fig. 22-12b. The coupled inductor is represented by an inductor on the primary and an ideal transformer between the primary and secondary. The coupling coefficient is assumed to be 1. Assuming the primary to secondary turns ratio is $1 : n$, we get

$$\bar{v}_L = V_{in}DT_s - \frac{V_o}{n}(1 - D)T_s = 0 \quad (22-21)$$

$$\Rightarrow \frac{V_o}{V_{in}} = n \frac{D}{1 - D} \quad (22-22)$$

$$\Rightarrow \frac{I_{in}}{I_o} = n \frac{D}{1 - D} \quad (22-23)$$

Although the analysis presented here assumes that the inductor current never goes to zero (called *continuous conduction mode* or CCM), it is very common to design flyback converters so that the inductor current does go to zero within each switching cycle. This operation, known as *discontinuous conduction mode* (DCM), leads to simplification of control design for flyback converters [5]. It should be noted that requirement of electrical isolation is not the only reason that a transformer or (coupled inductor) is used. Another important reason is that the transformer turns ratio leads to better utilization of power semiconductor devices.

22.3.4 Full-Bridge DC-DC Converter

Figure 22-13a shows the full-bridge dc-dc converter which is derived from the buck converter. The bridge circuit formed by switches S_1 , S_2 , S_3 , and S_4 converts the input dc voltage to a high-frequency ac (≥ 100 kHz), which is applied to the primary of transformer T_1 . The high frequency results in a small size for the transformer. After isolation, the high-frequency ac at the secondary of the transformer is rectified by the center-tapped diode bridge rectifier formed by D_1 and D_2 , and subsequently filtered by the L and C as in a buck converter. The topology is very popular for power levels greater than 500 W, when isolation is required.

Steady-state operating waveforms for the converter are shown in Fig. 22-13b. With switches S_3 and S_4 off, S_1 and S_2 are turned on simultaneously for $T_{on} = DT_s/2$, thereby applying a positive voltage across the transformer primary $T_{1,prim}$ and secondary $T_{1,sec1}$. During this time, diode D_1 conducts and a positive voltage appears across the inductor. With all the switches off, the transformer primary and secondary voltages are zero and the inductor current splits equally between diodes D_1 and D_2 . In the second half of the switching cycle S_1 and S_2 are off, while S_3 and S_4 are simultaneously turned on $DT_s/2$. The rectified voltage waveform is similar to that in the buck converter and is at double the switching frequency of the each switch. The magnetizing flux in the transformer is bidirectional (Fig. 22-13b), resulting in better utilization of the core as discussed in *Transformers Design*. The conversion ratio is similar to the buck converter, but scaled by the secondary to primary transformer turns ratio.

22.3.5 Other Isolated DC-DC Converters

Several other isolated converters are based on the buck converter. Figure 22-14a shows the forward converter. The operation and conversion ratio is similar to the buck converter. However, the output voltage is scaled by the transformer turns ratio, an additional winding and diode (D_R) are needed to reduce the core flux to zero in each switching cycle, and an additional diode (D_2) is required at the output. The forward and flyback converters have unidirectional core flux and are limited to low-power applications. The push-pull converter (Fig. 22-14b), also derived from the buck converter, is better suited for higher power levels, limited by voltage rating of the switches required. Details of these converters can be found in Refs. [4, 5].

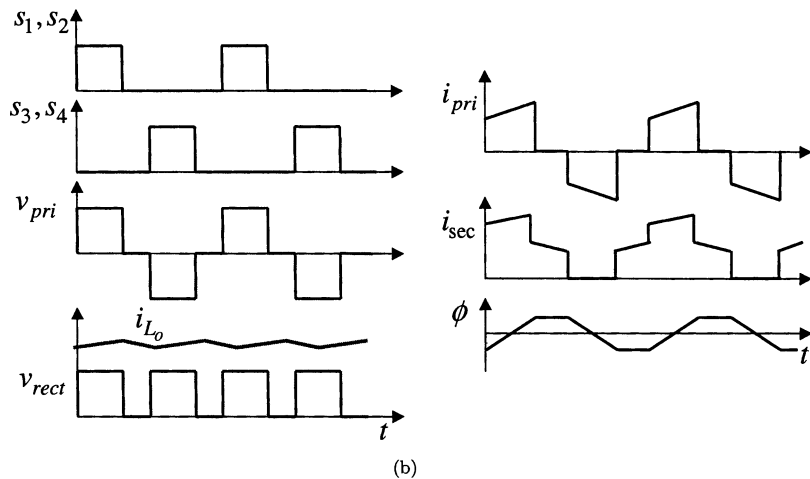
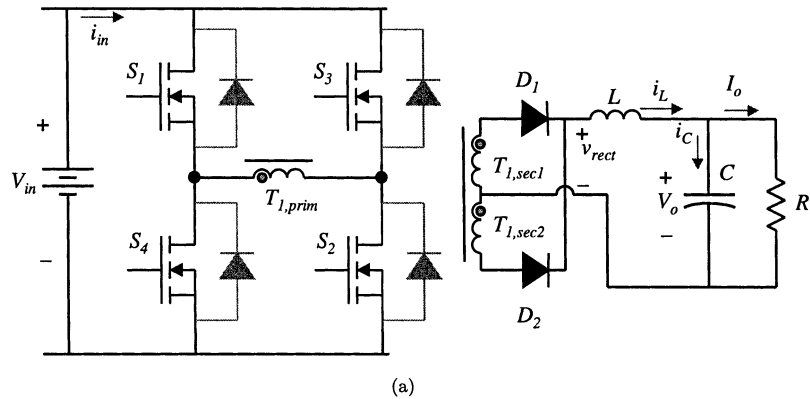


FIGURE 22-13 Full bridge dc-dc converter: (a) circuit, (b) steady-state waveforms.

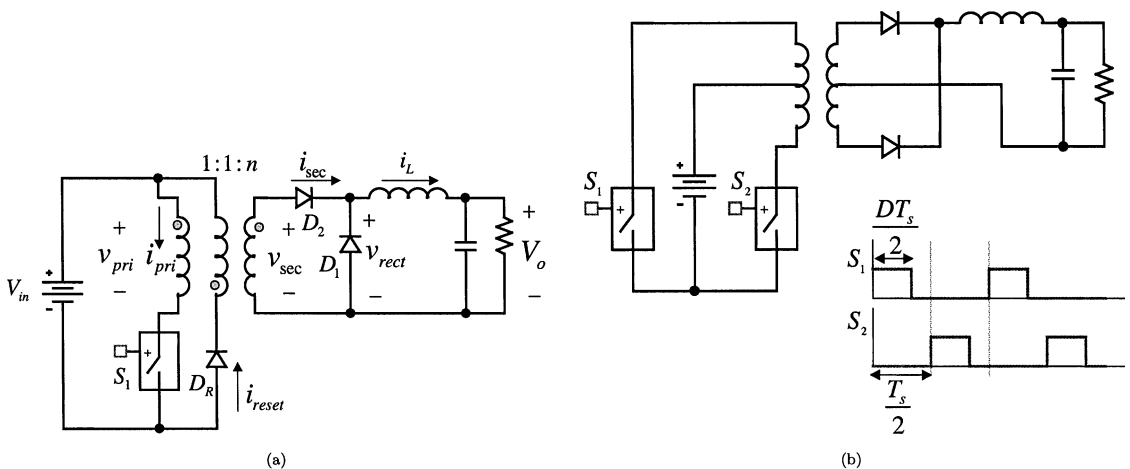


FIGURE 22-14 Other isolated converters: (a) Forward converter, (b) Push-Pull Converter.

22.3.6 Recent Developments and Future Trends

The size of filter components (inductor and capacitor) and isolation transformer reduce as the switching frequency is increased. Thus, a high switching frequency is desirable to minimize size and weight. However, parasitics and other nonidealities in dc-dc converters eventually limit the switching frequency and efficiency. For example, in flyback and forward converters, leakage inductance of the coupled inductor/transformer is a significant problem at high frequencies; in each switching cycle all the energy stored in the leakage inductance at switch turn-off has to be dissipated. Similarly, during turn-on of switches energy stored in parasitic output capacitance of the switch is dissipated inside the switch. These losses increase in proportion to the switching frequency. Thus, thermal or efficiency consideration eventually limit the maximum switching frequency. Besides the ones mentioned above other limiting factors are: switching times of diodes and MOSFETs, reverse recovery of diodes, capacitance of schottky diodes, and capacitance of transformers. To overcome these limitations, several circuits have been developed, which utilize the parasitic inductance and capacitances to advantage. Although the modifications in these sometimes add disadvantages, for specific applications, the advantages outweigh the disadvantages.

Soft-switching converters use resonance conditions between parasitic capacitance and inductance so that either the capacitance of switching devices is discharged before the device is actually turned on, or the current through the leakage inductance is reduced to zero prior to turning off. In some circuits, additional inductors and/or capacitors are added to produce resonance conditions. These converters, generically called *resonant converters*, are widely used in applications such as computer power supplies, electronic ballasts for fluorescent lamps, battery charging, and various portable applications. Details of these converters can be found in Refs. [4, 5].

Reduction in filtering requirements has also been achieved by using *interleaving*. An interleaved converter or multiphase converter has two or more converters called *phases*. These phases operate in parallel with their inputs and outputs being common. They are switched out of phase (180° for a 2-phase case, 120° for three phases, etc.) so that the ripple currents in the individual inductors are also out of phase. This results in lower effective current ripple both at the output and the input, and thus, smaller filter size for a given ripple specification. The lower values of the inductor also lead to faster dynamic response. Hybrid converters combining the benefits of soft switching with lower filter requirements have also been developed [13, 14].

To reduce size and minimize the number of discrete components, there is a significant effort in integrating all the semiconductors in one package. For example, on semiconductors [15] and power integration [16] have developed modules for use in off-line flyback converters; converters whose input is rectified line voltage are called off-line converters. The module contains a high-voltage power MOSFET and control circuit in one standard package. Similar modules are also available for low-power dc-dc converters (e.g., see Ref. [17]). Efforts are also being made to integrate all the magnetic components in dc-dc converters, using one single magnetic component, a concept called integrated magnetics.

22.4 FEEDBACK CONTROL OF POWER ELECTRONIC CONVERTERS

In the last section we saw that the steady-state output of a dc-dc converter, usually the output voltage, is controlled by the duty ratio. To account for changes in load current, input voltage, losses, and nonidealities in the converter, feedback based automatic control is required. Figure 22-15 shows a block diagram of output voltage control for a buck converter. The laplace domain control block diagram is also shown. The sensed output voltage is multiplied by a feedback gain $G_{FB}(s)$ before being compared with a reference value. The error is fed to an appropriate error compensator that generates a control voltage v_c , which is converted to duty ratio d by the PWM block.

Toward designing a suitable controller, we will first describe a dynamic model of the power converter and then a simple loop-shaping control design method based on input to output bode plots.

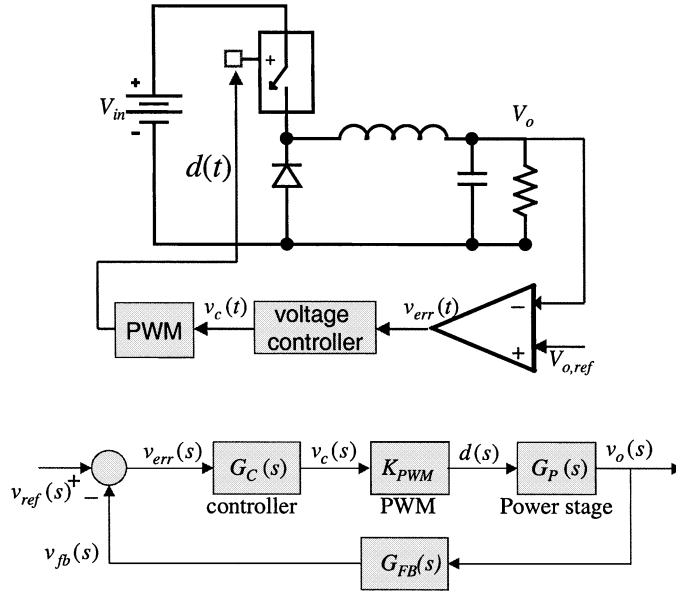


FIGURE 22-15 Block diagram of output voltage control for a buck converter.

It is possible to design more complex controllers in order to meet specific requirements, and the converter topology or operating method may also be modified to make the control design easier (e.g., DCM operation of flyback converters). To keep the explanation simple, it is assumed that the converter operates in CCM.

22.4.1 Dynamic Modeling

The power converter essentially consists of the PWM block and the power stage itself. The feedback gain G_{FB} is usually a constant. The PWM block shown in Fig. 22-10 converts the input control voltage $\bar{v}_c(t)$ to a duty ratio $d(t)$. From geometrical considerations

$$d(t)/\bar{v}_c(t) = 1/\hat{V}_{ramp} = K_R \quad (22-24)$$

where $\hat{V}_{ramp}$ is the peak value of the ramp $v_{ramp}(t)$. The power stage transfer function from $d(s)$ to $\bar{v}_o(s)$ can be derived using one of the methods stated below.

Dynamics of Averaged Quantities. As stated earlier, the bipositional switch approach and averaging are valid for analyzing low-frequency dynamics ($< f_s/5$) of the power converter. Unlike steady-state analysis, under dynamic conditions $\bar{v}_L \neq 0$ and $\dot{\bar{i}}_C \neq 0$. Averaging the instantaneous state equations [Eq. (22-15)] over one switching cycle, dynamics of the averaged inductor current and capacitor voltage in a buck converter are

$$L \dot{\bar{i}}_L = d(t) \cdot V_{in} - \bar{v}_o \quad (22-25)$$

$$C \dot{\bar{v}}_o = \bar{i}_L - \bar{v}_o/R_L \quad (22-26)$$

Here the time varying duty cycle $d(t)$ is the control input and the averaged output voltage $\bar{v}_o$ is the output that has to be regulated. The situation for the buck converter is simple because the model described by Eqs. (22-25) and (22-26) is linear if V_{in} and R_L are assumed constant, for which case

exact transfer functions describing large signal behavior can be derived. For boost and buck-boost converters, the averaged state equations involve terms with multiplications of $d(t)$ with a state variable. Thus, the model has to be linearized, and small signal dynamics obtained at different operating points are utilized for linear control design. It is of course possible to design large signal control based on the nonlinear model at the expense of mathematical complexity [18]. However, ease of design and simple cost-effective implementation has made linear design the preferred method in most power electronic converters in the low-to-medium power range.

Averaged Circuit Representation. Instead of writing averaged state equations explicitly as in Eqs. (22-25) and (22-26), an averaged circuit representation of the bipositional switch can be derived and substituted in different converter circuits Refs. [19–21]. As shown in Fig. 22-16a the bipositional switch can be considered as a two-port network with a voltage port (subscript_{vp}) at the input and a current port (subscript_{cp}) at the output. The average voltage and currents of the two ports are related as

$$\bar{v}_{cp}(t) = d(t) \cdot \bar{v}_{vp}(t) \quad (22-27)$$

$$\bar{i}_{vp}(t) = d(t) \cdot \bar{i}_{cp}(t) \quad (22-28)$$

The relations in the above equations correspond to those of an ideal transformer with turns ratio of $1 : d(t)$. Thus, for analysis purposes, the bipositional switch can be modeled as an ideal transformer whose turns ratio $d(t)$ can be controlled as shown in Fig. 22-16b. This representation is extremely useful in conjunction with circuit simulators which can perform operating point (dc bias) calculations, linearization, and ac analysis. Parasitic effects, like series resistances of inductors and capacitors,

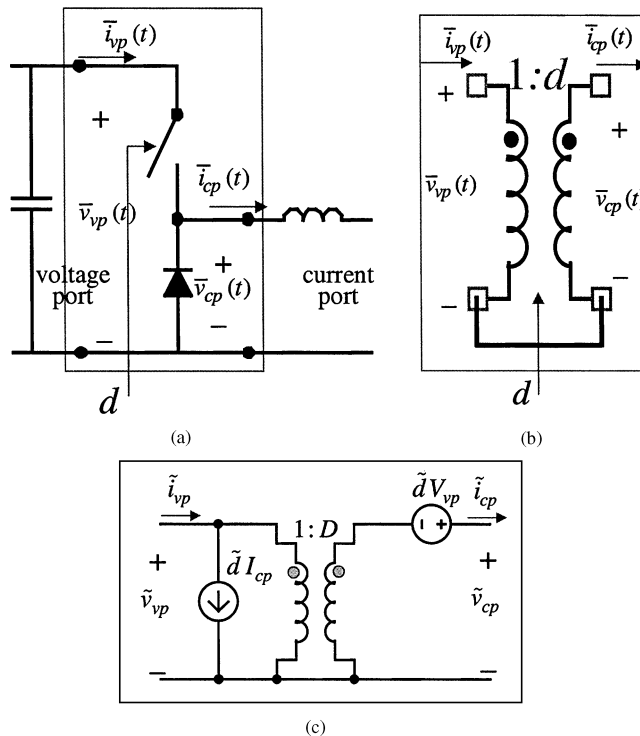


FIGURE 22-16 Bipositional switch: (a) two port network, (b) average representation, (c) small signal model.

can be easily incorporated in the averaged circuit model. Circuit simulators like SPICE [22], Saber, and Simplorer are commonly used for this purpose. The small-signal circuit representation of the averaged circuit obtained via linearization is shown in Fig. 22-16c. Quantities in upper case indicate operating point values, while the quantities with a “~” indicate small signal perturbations about the operating point. This representation can be utilized to derive small signal transfer functions using circuit analysis techniques.

22.4.2 Control Design

For a dc-dc converter, the main control objectives are: stability, zero steady-state error, specified transient response to step change in reference and in disturbance inputs (load and input voltage), and robustness to parametric changes. Transfer functions of different components— K_R , $G_{PS}(s)$, and $G_{FB}(s)$ —are obtained as described above. The error compensator is then designed so that the open loop transfer function $G_{OL}(s)$ has a specified gain crossover frequency and phase margin. The gain cross-over frequency determines the response time of the controlled converter to changes in reference voltage and load current. Phase margin is usually in the range of 45° to 60° depending on the overshoot tolerable. Details on relation between gain crossover frequency, phase margin, and transient response can be found in any textbook on linear control (e.g., see Ref. 23).

An integrator (pole at origin) is added in $G_c(s)$ to obtain zero steady-state error. Zeroes are added at appropriate locations to obtain required phase margin. The dc gain of $G_c(s)$ is adjusted to achieve the required crossover frequency. Finally, to improve noise immunity, poles may be added for fast roll-off of the gain after the cross-over frequency. A systematic loop-shaping procedure along with implementation suited to dc-dc converters is described in Ref. [24].

Example: Voltage Control of a Buck Converter. A controller has to be designed to regulate the output voltage of a buck converter to a constant value. The specifications, parameters, and control requirements are listed in Table 22-2. Using the methods described above, the duty ratio to output voltage transfer function can be derived to be

$$\frac{\tilde{v}_o(s)}{\tilde{d}(s)} = \frac{V_m(1 + sCR_{ESR})}{1 + s[CR_{ESR} + L/R_L] + s^2LC[1 + R_{ESR}/R_L]} \quad (22-29)$$

The transfer function has a complex pole pair due the L-C filter, and a left half zero due to the ESR of the output capacitor. The compensator designed in accordance with the aforementioned considerations is

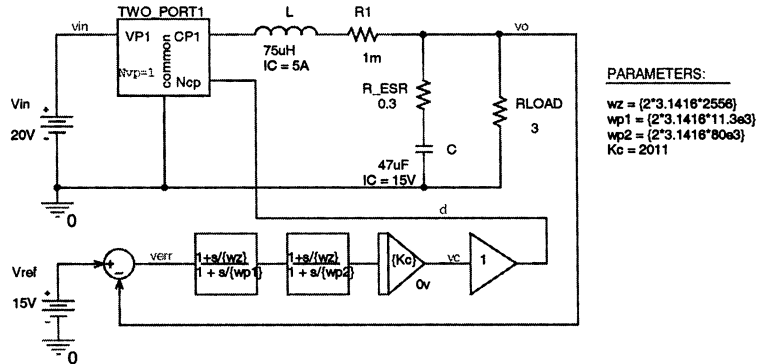
$$G_c(s) = \frac{K_c(1 + s/\omega_z)^2}{s(1 + s/\omega_{p1})(1 + s/\omega_{p2})} \quad (22-30)$$

where $K_c = 2011$, $\omega_z = 2\pi \times 2556$ rad/s, $\omega_{p1} = 2\pi \times 11.3e3$ rad/s, and $\omega_{p2} = 2\pi \times 80e3$ rad/s.

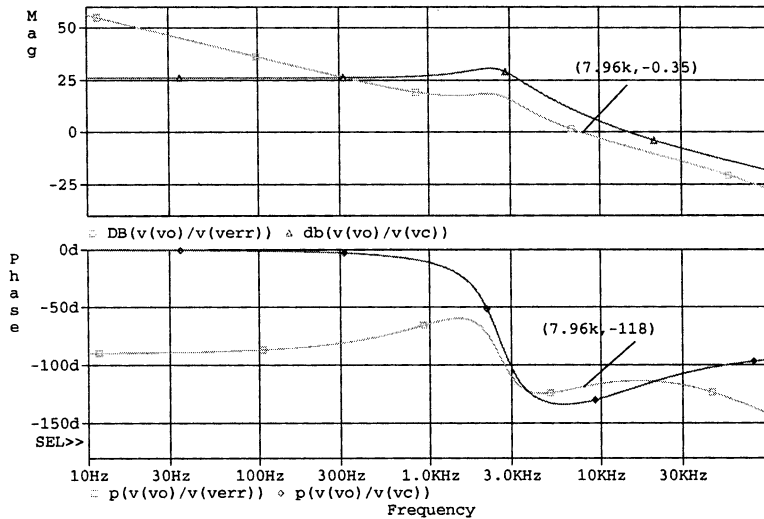
Representation of the controlled converter using ORCAD PSpice [25] is shown in Fig. 22-17a. The bipositional switch has been replaced by a two-port network that models a transformer with

TABLE 22-2 Control Design Example

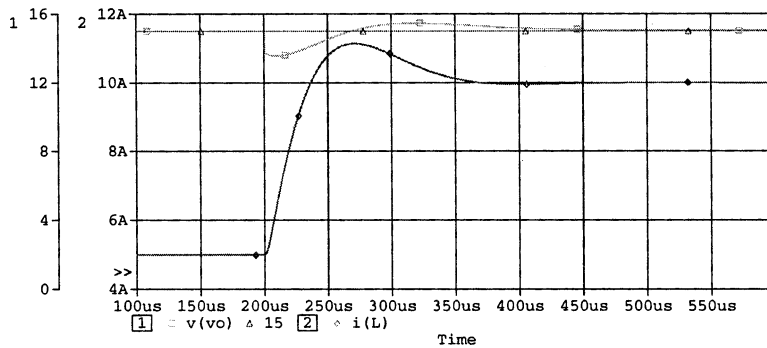
Specifications		Parameters		Requirements	
Input voltage	20 [V] to 30 [V] dc	L	75 [μ H]	Cross-over frequency	8 [kHz]
Output voltage	15 [V] dc	C	47 [μ F]	Phase margin	$> 60^\circ$
Maximum load current	5 [A]	R_{ESR}	0.3 [Ω]		
Switching frequency	200 [kHz]	K_R	1		



(a)



(b)



(c)

FIGURE 22-17 Controlled buck converter: (a) averaged representation using PSpice, (b) open loop bode plots, (c) transient response.

controllable turns ratio. The compensated and uncompensated transfer functions obtained using ac analysis are shown in Fig. 22-17b. The gain crossover frequency and the corresponding phase of the compensated transfer function are indicated. Figure 22-17c shows dynamic response of the controlled converter when a step change in load is applied at 0.2 ms.

22.4.3 Current Mode Control

In most converters, the inductor current is an *internal* state of the power converter. Changes in the input voltage and duty ratio are first reflected in the inductor current, and subsequently in the output voltage. Thus, controlling the inductor current can lead to better performance. Figure 22-18 shows a cascaded control structure where the internal current controller is about an order of magnitude faster than the outer voltage loop. The average value of the inductor current is controlled to a reference that is generated by the error compensator for the voltage-control loop. The current controller is designed using the transfer function from the duty ratio to the inductor current. For voltage controller, the current control loop is assumed to be ideal, that is, $i_L = i_{L,ref}$; this is justified since the current controller is much faster than the voltage controller. The voltage compensator is then designed, using the inductor current to the output voltage transfer function. In a buck or buck-derived topology, the average inductor current is equal to the load current. Thus, fast control over the inductor current effectively mitigates steady state and transient variations in the input voltage without affecting the output voltage. This current control method is called *average current control*.

Another popular method is *peak current mode control*. In this method, the peak value of the inductor current is controlled to the reference value (generated by the voltage-control loop) in each switching cycle. Peak current mode control has the additional advantage of balancing the positive and negative flux excursions in transformer isolated topologies like full bridge and push pull. However, peak current mode control requires extra precautions to avoid subharmonic and chaotic operation [4, 5, 26].

22.4.4 Other Control Techniques

The basic modeling method described above is applicable to other types of converters (like dc-ac) as long as low-frequency behavior is being studied. Dynamics of the filter elements may of course be different. In dc-ac converters, the control objective is usually to track a moving reference (e.g., sinusoidal control voltage or current). Using a stationary to rotating frame transformation, commonly called the *abc to dq transformation*, the tracking problem can often be reduced to a regulation problem. If current control is implemented in the stationary frame, where the control objective is to track a sinusoidally varying reference, then either average current control or *hysteretic current control* is used. In hysteretic current control, the current is maintained in a band about the reference

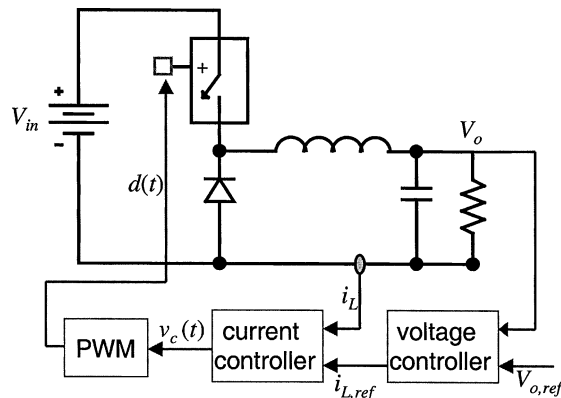


FIGURE 22-18 Average current control of a buck converter.

value. If the current error is below the lower limit of the band, a positive voltage is applied across the inductor (switch on in a buck converter) to increase the current. To reduce the current a negative or zero voltage is applied across the inductor. With hysteretic control, the rise and fall times are only limited by the power components of the converter. However, it has the disadvantage of variable switching frequency, whose instantaneous value depends on a combination of several factors. Several other techniques have recently been proposed for control of power electronic converters: sigma-delta, sliding mode, dead beat, etc. Details of these control techniques can be found in Ref. [27]. In digital implementations, predictive current control is commonly used to reduce the effect of sensing and computational delays. So far, digital control is only used in high power converters, where the overall cost justifies the cost of digital and interface components. However, there is significant effort in extending the benefits of digital control to lower power converters.

22.5 DC-AC CONVERSION: INVERSION

DC-ac converters constitute a significant portion of power electronic converters. These converters, also called *inverters*, are used in applications such as electric motor drives, uninterruptible power supplies (UPS), and utility applications such as grid connection of renewable energy sources. Inverters for single phase ac and 3-phase 3-wire ac systems (without a neutral connection) are described in this section.

22.5.1 Single Phase AC Synthesis

In an ac system both the voltage and the current should be able to reverse in polarity. Further, the voltage and current polarities may or may not be the same at a given time. Thus, a dc-ac converter implementation should be able to output a voltage independent of current polarity. In the full-bridge dc-dc converter shown in Fig. 22-19a, the primary circuit consisting of four controlled switches, also called *H-bridge*, has two bipositional switch implementations. Each bipositional switch has bidirectional current capability but only positive output voltage ($v_{AN} > 0$, $v_{BN} > 0$). However, based on the duty cycles, the difference of the outputs, $V_{AB} = v_{AN} - v_{BN}$, can reverse in polarity. Thus, the H-bridge is used for synthesizing single phase ac voltage from a dc voltage.

Quasi-square Wave Inverter. The simplest form of dc-ac conversion, albeit with poor quality, is synthesis of quasi-square wave ac instead of a pure sine wave. Diagonally opposite switches in the H-bridge are turned on simultaneously. The pulse width of each pair is controlled to adjust the magnitude of the fundamental component, while the switching frequency is equal to the required output frequency. The synthesized voltage waveform is shown in Fig. 22-19b. The peak value of fundamental and harmonic components are

$$V_{AB,n} = \frac{4V_{in}}{n\pi} \sin(n\pi d/2) \quad n \text{ odd} \quad (22-31)$$

where d is the duty ratio and n is the harmonic number. This converter is widely used for low cost low power UPS applications where the voltage waveform quality is not important. Incandescent lighting, universal input motors, and loads with a diode bridge or power factor corrected front end (discussed in Sec. 22.8.1) are not affected by the voltage waveform quality. The load current, i_{AB} , has harmonics based on the load characteristics. Sometimes an LC filter is added at the output to reduce the voltage (and therefore the current) harmonics.

Single-Phase Sinusoidal Voltage Synthesis. For applications requiring low voltage and current distortion, high-frequency PWM is utilized to generate a sinusoidally varying average voltage. The power converter used is the H-bridge shown in Fig. 22-19a. The duty ratio for each bipositional switch, also called one leg of the inverter, is varied sinusoidally. The switching signals are generated

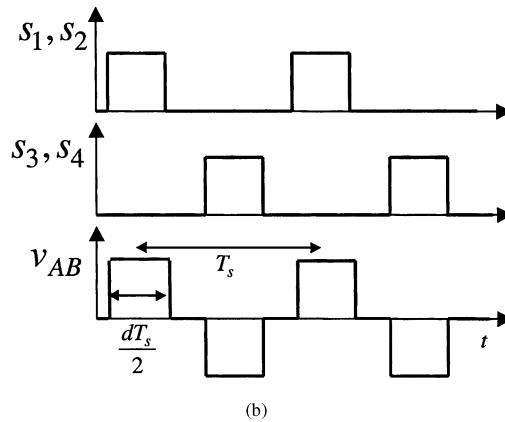
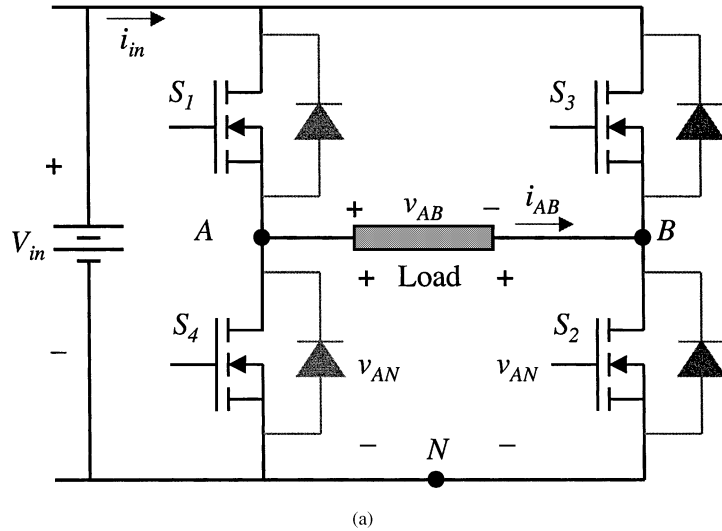


FIGURE 22-19 Single-phase inverter: (a) circuit, (b) quasi-square wave synthesis.

by comparison of a sinusoidally varying control voltage with a *triangle* wave as shown in Fig. 22-20. Equations relating the control voltages, duty ratios, and the averaged output voltages are as follows:

$$v_c = \hat{V}_c \cdot \sin(\omega_m t) \quad (22-32)$$

$$v_{cA}(t) = v_c = \hat{V}_c \cdot \sin(\omega_m t) \quad (22-33)$$

$$v_{cB}(t) = -v_c = -\hat{V}_c \cdot \sin(\omega_m t) \quad (22-34)$$

$$d_A(t) = \frac{1}{2} \left(1 + \frac{v_{cA}}{\hat{V}_{tri}} \right) = \frac{1}{2} (1 + m \cdot \sin(\omega_m t)) \quad (22-35)$$

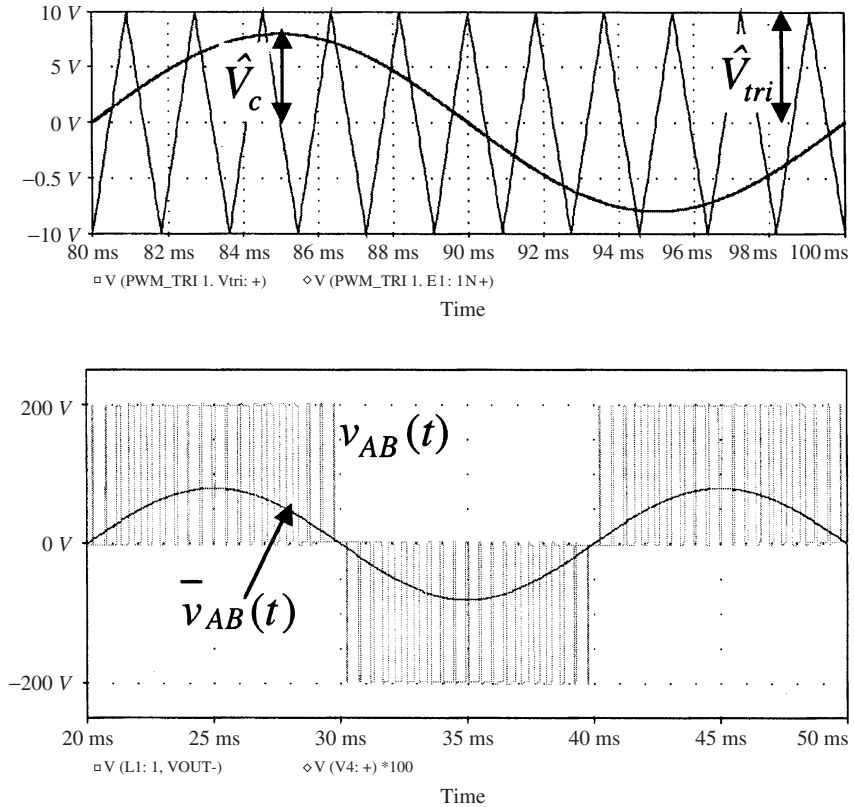


FIGURE 22-20 Single-phase sinusoidal ac synthesis waveforms.

$$d_B(t) = \frac{1}{2} \left(1 + \frac{v_{cB}}{\hat{V}_{tri}} \right) = \frac{1}{2} (1 - m \cdot \sin(\omega_m t)) \quad (22-36)$$

$$\bar{v}_{AN}(t) = d_A(t) \cdot V_{in} = \frac{1}{2} (1 + m \cdot \sin(\omega_m t)) \cdot V_{in} \quad (22-37)$$

$$\bar{v}_{BN}(t) = d_B(t) \cdot V_{in} = \frac{1}{2} (1 - m \cdot \sin(\omega_m t)) \cdot V_{in} \quad (22-38)$$

$$\bar{v}_{AB}(t) = (d_A(t) - d_B(t)) \cdot V_{in} = m \cdot V_{in} \cdot \sin(\omega_m t) = \underbrace{(V_{in}/\hat{V}_{tri})}_{k_{PWM}} \cdot v_c(t) \quad (22-39)$$

Here $\hat{V}_c$ and V_{tri} are peak values of control voltage and the triangle wave, respectively, $m = \hat{V}_c/\hat{V}_{tri} \in [0, 1]$ is the modulation index, $\omega_m = 2\pi f_m$ is the angular frequency of the sinusoid to be synthesized, while $d_A(t)$ and $d_B(t)$ are duty ratios of switches S1 and S3, respectively. In Eq. (22-39) k_{PWM} may be regarded as the gain of the power converter that amplifies the control signal $v_c(t)$ to the average output voltage $\bar{v}_{AB}(t)$. The maximum peak value of the output voltage, obtained for $m = 1$, is V_{in} . This is significantly lower than that obtainable with the quasi square wave inverter ($4V_{in}/\pi$). However, harmonics

in the output voltage are significantly reduced and are at much higher frequencies: $k \cdot f_s \pm l \cdot f_m$, where k and l are integers such that $k + l$ is odd [4]. The switching frequency $f_s \geq 20$ kHz is significantly higher than the output frequency f_m , which usually has a maximum value of about 50/60 Hz. If the load is inductive, the current harmonics are reduced further, and the current is almost sinusoidal.

Equation (22-37) can be rewritten as

$$\bar{v}_{AN}(t) = d_A(t) \cdot V_{in} = \frac{V_{in}}{2} + \frac{k_{PWM}}{2} \cdot v_c(t) \quad (22-40)$$

This clearly shows that on an average basis the “neutral point” for the output of one inverter leg is $V_{in}/2$ above “N,” that is, at the mid-point of the input dc bus. Thus, using the same H-bridge a split-phase ac (two ac voltages 180° out of phase with a common return) can be generated if the center point of the dc bus is available as the neutral connection for the output.

22.5.2 Three-Phase AC Synthesis

The last observation in the previous section leads us to 3-phase inverters without a neutral connection. The circuit consists of three legs, one for each output with a common dc link as shown in Fig. 22-21a.

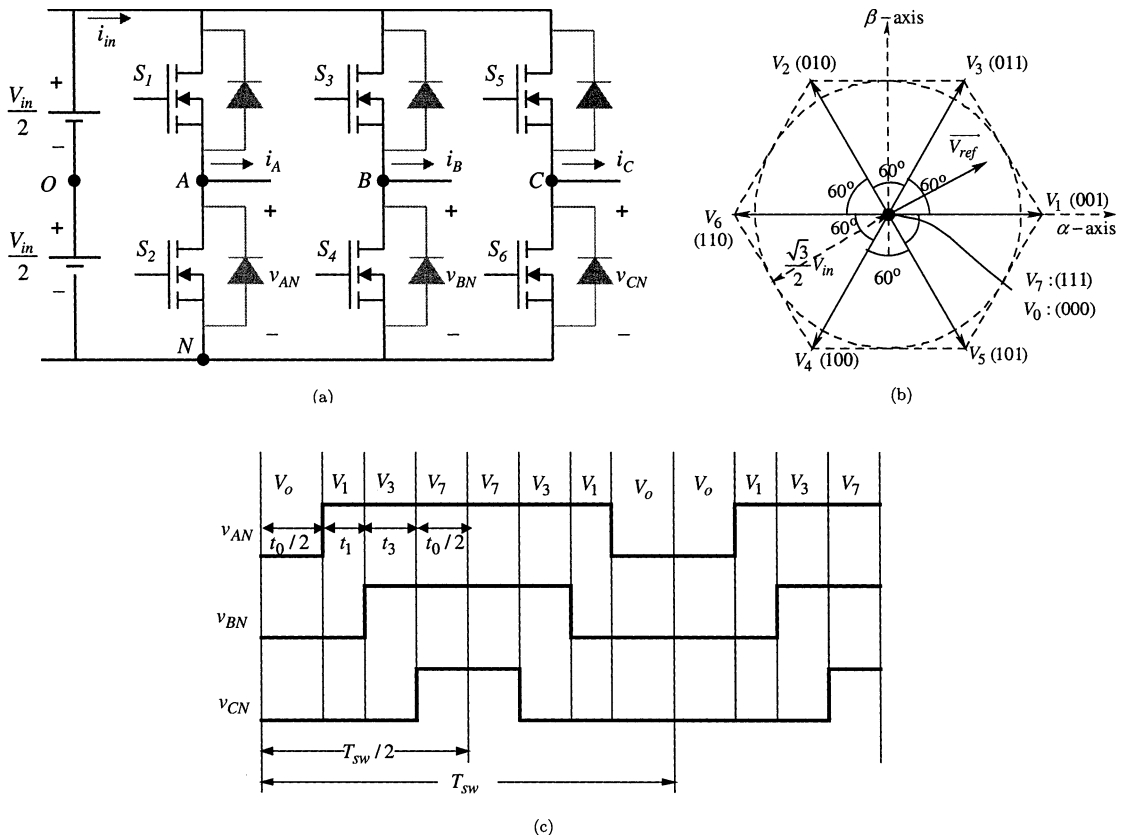


FIGURE 22-21 3-Phase ac synthesis: (a) converter, (b) output voltage vectors, (c) instantaneous waveforms.

Using sine triangle PWM with control voltages offset by 120° (instead of 180° as in the single phase case) we obtain

$$v_{cA}(t) = \hat{V}_c \cdot \sin(\omega_m t) \quad (22-41)$$

$$v_{cB}(t) = \hat{V}_c \cdot \sin(\omega_m t - 2\pi/3) \quad (22-42)$$

$$v_{cC}(t) = \hat{V}_c \cdot \sin(\omega_m t + 2\pi/3) \quad (22-43)$$

$$\bar{v}_{AN}(t) = \frac{V_{in}}{2} + \frac{k_{PWM}}{2} \cdot v_{cA}(t) \quad (22-44)$$

$$\bar{v}_{BN}(t) = \frac{V_{in}}{2} + \frac{k_{PWM}}{2} \cdot v_{cB}(t) \quad (22-45)$$

$$\bar{v}_{CN}(t) = \frac{V_{in}}{2} + \frac{k_{PWM}}{2} \cdot v_{cC}(t) \quad (22-46)$$

The zero-sequence component of the output voltages, $v_z = (v_{AN} + v_{BN} + v_{CN})/3 = V_{in}/2$, does not appear in the line-to-line voltages, and since there is no neutral connection to the inverter zero-sequence currents do not flow.

The maximum peak value of the output line-to-line voltages is $\hat{V}_{LL} = (\sqrt{3}/2)V_{in}$. Using square wave inversion, similar to that for the single-phase case, we can obtain higher magnitude for the fundamental component of the output voltages at the cost of adding harmonics. However, if, instead of all the harmonics, only the fundamental and those harmonics of the square wave that contribute zero-sequence component (triplen harmonics) are retained, the output voltage amplitude increases without adding harmonics to the line-to-line voltages and the line currents. Usually, addition of the third harmonic component is sufficient Refs. [28, 29]. As described in Refs. [30, 31], the most effective method is to add the following zero-sequence component to the control voltages for each phase

$$v_{cz}(t) = \frac{1}{2} [\max(v_{cA}(t), v_{cB}(t), v_{cC}(t)) + \min(v_{cA}(t), v_{cB}(t), v_{cC}(t))] \quad (22-47)$$

In terms of output voltage generation, this is equivalent to space vector modulation (SVM).

22.5.3 Space Vector Modulation

This method has become extremely popular for 3-phase inverters in the low-to-medium power range. A very brief description will be presented here and details can be found in Refs. [27, 28, 30].

For 3-phase systems with no zero-sequence component, that is, $v_z = (v_{AN} + v_{BN} + v_{CN})/3 = 0$, the 3-phase quantities are linearly dependent and can be transformed to a 2-phase orthogonal system commonly called the $\alpha\beta$ system. Quantities in the $\alpha\beta$ system can be represented by complex numbers and as two-dimensional vectors in a plane, called *space vectors*. The transformation from the abc to $\alpha\beta$ quantities is given by

$$\vec{v}_{\alpha\beta}(t) = v_\alpha(t) + j \cdot v_\beta(t) = e^{j0} \cdot v_\alpha(t) + e^{j2\pi/3} \cdot v_\beta(t) + e^{j4\pi/3} \cdot v_\gamma(t) \quad (22-48)$$

With negative sequence components absent, α and β components of steady-state sinusoidal abc quantities are also sinusoids with constant amplitude and a 90° phase difference between them. Under transient conditions they are arbitrary time-varying quantities. Thus, for balanced sinusoidal conditions, the space vector $\vec{v}_{\alpha\beta}(t)$ rotates in counter clockwise direction with angular frequency equal to frequency of the abc voltages, and describes a circle of radius $(3/2)\hat{V}_{ph}$, $\hat{V}_{ph}$ being the peak of the phase voltage.

The instantaneous output voltages of the 3-phase inverter shown in Fig. 22-21a can assume eight different combinations based on which of the six MOSFETs are on. The space vectors for these eight combinations are shown in Fig. 22-21b. For example, vector V_4 denoted by (100) corresponds to switch states $v_{AN} = V_{in}$, $v_{BN} = 0$, and $v_{CN} = 0$. The vectors $V_0(000)$ and $V_7(111)$ have zero magnitude and are called *zero vectors*.

Synthesis utilizing the idea of space vectors is done by dividing one switching time period into several time intervals, for each of which a particular voltage vector is the output by the inverter. These time intervals and the vectors applied are chosen so that the average over one switching time period is equal to the desired output voltage vector. For the reference voltage vector $\vec{v}_{ref}$, shown in Fig. 22-21b, the nonzero vectors adjacent to it (V_1 and V_3), and the zero vectors (V_0 and V_7) are utilized as shown in Fig. 22-21c. Relative values of time intervals t_1 and t_3 determine the direction, while ratio of t_0 to the switching time period determines the magnitude of the output vector synthesized.

The maximum obtainable average vector lies along the hexagon connecting the six nonzero vectors. As stated earlier, balanced 3-phase sinusoidal quantities describe a circle in the $\alpha\beta$ plane. Thus, to synthesize distortion-free and balanced 3-phase sinusoidal voltages the circle must be contained within the hexagon, that is, with a maximum radius of $\sqrt{3}/2 \cdot V_{in}$. This gives the maximum peak value of line-to-line voltage obtained with SVM as $\hat{V}_{LL} = V_{in}$. This is significantly higher than that obtained using sine triangle PWM: $\sqrt{3}/2 \cdot V_{in}$. Further, the sequence and choice of vectors applied can be optimized to minimize number of switchings and ripple in the resulting currents [32]. There are several variations of SVM, each suited to a different application. Space vector modulation can be easily implemented digitally using microcontroller and digital signal processor (DSPs), and is extremely advantageous in control of 3-phase ac machines strategies using *vector control* and *direct torque control* (DTC) [33–37].

22.5.4 Multilevel Converters

The converter topologies described so far are based on a 2-level converter leg (bipositional switch), where the output voltage of each leg (v_{AN}) can be either zero or V_{in} . The converters are therefore called 2-level converters. In 2-level converters, all the switches have to block the full dc bus voltage (V_{in}). For high-power applications insulated gate bipolar transistors (IGBTs) and gate turn-offs (GTOs) are used as the semiconductor switches. These have higher voltage and current ratings, and lower on-state voltage drop compared to power MOSFETs, but cannot switch as fast. In some applications like some motor drives and utility applications, even the voltage ratings of available IGBTs and GTOs is not sufficiently high. Simple series connection, to achieve a higher blocking voltage, has problems of steady state and dynamic voltage sharing. Moreover, due to the low switching frequency of high-power switches, the output voltage and current quality deteriorates. These issues are addressed by multilevel converters. In a multilevel converter [38, 39], the output of each phase leg can attain more than two levels leading to improved quality of the output voltage and current. The circuit comprising each leg and its proper operation ensure that voltage blocked by the switches reduces as the number of levels is increased. In addition, multilevel converters are modular to some extent, thereby making it easy to scale voltage ratings by increasing the number of “cells”.

Multilevel PWM. For 2-level PWM, comparison of the control voltage with a triangle wave generates the switching signal for the top switch, while the bottom switch is controlled in complement to the top switch. Each of these two states corresponds to the two levels of the output voltage. For multilevel converters, there are more than two effective switch states, each of which corresponds to an output voltage level. For example, in a 3-level converter there are three effective states $q(t) = 0, 1, 2$, corresponding to output voltage levels $v_{AN}(t) = 0, V_{in}/2, V_{in}$. The control voltage $v_c(t)$ is compared with two triangle waves to obtain two switching signals $q_1(t)$ and $q_2(t)$, and the effective switching signal can be obtained as $q(t) = q_1(t) + q_2(t)$ as shown in Fig. 22-22. The output voltage is then given by $v_{AN} = q(t) \cdot (V_{in}/2)$. Switching signals for the individual switches are derived using $q(t)$ and the circuit topology. For the waveforms in Fig. 22-22, $f_s = 60\text{Hz}$ and $V_{in} = 2\text{ kV}$. Since the

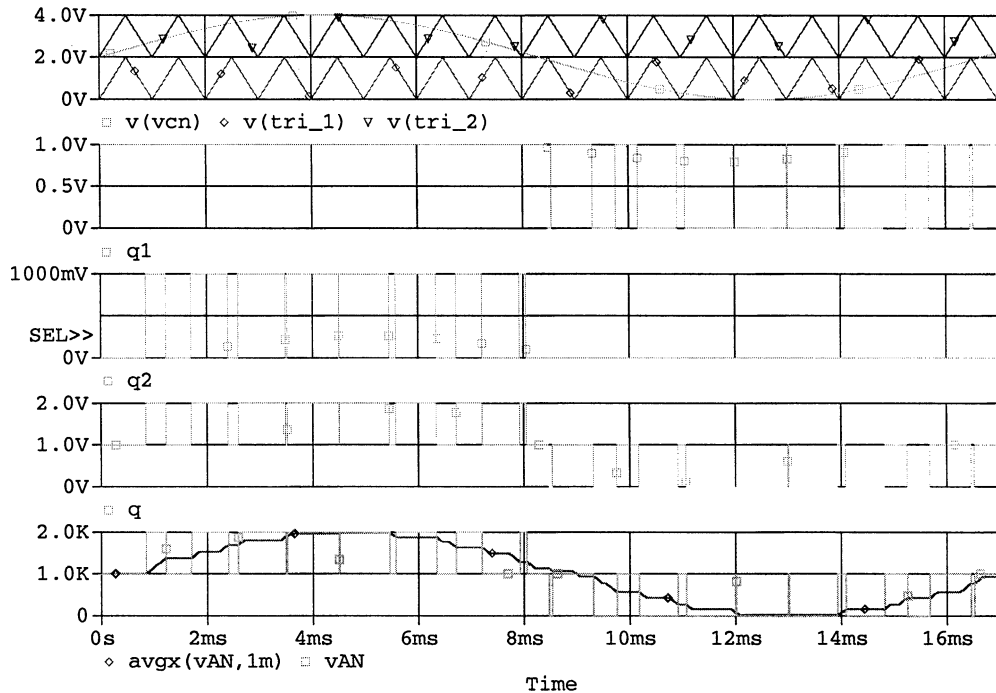


FIGURE 22-22 Multilevel triangle comparison.

v_{AN} waveform is closer to desired sinusoid in the 3-level case, the output voltage has lower total harmonic distortion (THD) even if the switching frequency is low. For 3-phase converters, space vector-based PWM can be used for generating the switching signals (e.g., see Ref. 40), the advantage in the multilevel case compared to the 2-level case being the significantly higher number of output voltage vectors.

Multilevel Converter Topologies. There are three basic multilevel converter topologies—diode clamped, flying capacitor, and cascaded full-bridge converters.

Diode Clamped Converter. Figure 22-23a shows 1-phase leg of a 3-level diode clamped converter [41]. The input dc bus is split by means of capacitors. Pairs of switches are turned on to obtain three different voltage levels for the output voltage $v_{AN} = 0, V_{in}/2, V_{in}$, as shown in Fig. 22-23c. It is evident that this circuit acts like a *tripositional* switch connecting the output to one of three positions of the input dc bus. The minimum voltage at point b_1 , and the maximum voltage at point b_2 , is clamped to $V_{in}/2$ by the blocking diodes D_{b1} and D_{b2} , respectively. Thus, all the switches have to block $V_{in}/2$ during their off state. This topology can be extended to more number of levels. However, it is eventually limited by the voltage rating of blocking diodes, which have to block increasing voltages as the number of levels is increased. One-phase leg of a 5-level version is shown in Fig. 22-23b.

Flying Capacitor Converter. Figure 22-24 shows the topology of a 3-level flying capacitor converter. The basic idea here is that the capacitor C is charged to half the input dc voltage by appropriate control of the switches. The capacitor can then be inserted in series with the output voltage—either adding or subtracting $V_{in}/2$, and thereby giving 3-output voltage levels.

Cascaded Full Bridge Converters. In this scheme [42], single-phase H-bridges shown in Fig. 22-19a are connected in series at the output to form one single phase circuit. Three separate

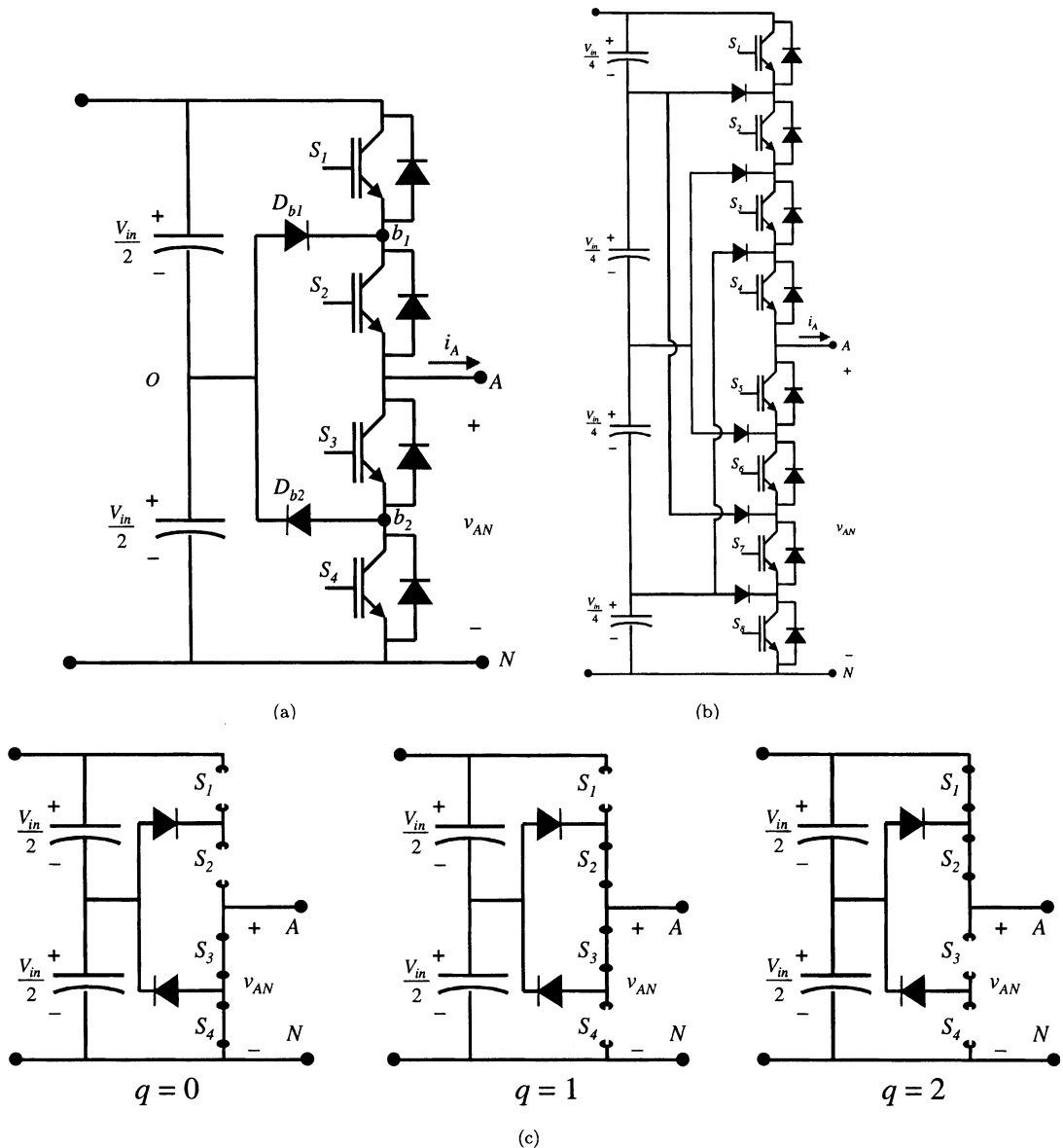


FIGURE 22-23 Diode clamped converters: (a) one phase of a 3-level converter, (b) one phase of 5-level converter, (c) switching states in a 3-level converter.

circuits are required for a 3-phase implementation. Since all the H-bridges are same, the circuit is modular and can be scaled by adding more H-bridges. However, dc sources at the input of all H-bridges have to be isolated from each other. It is also possible to combine different types of H-bridges—IGBT-based fast switching type and GTO-based slower switching type—or have different dc bus voltage magnitudes in different bridges to optimize losses or increase effective number of levels.

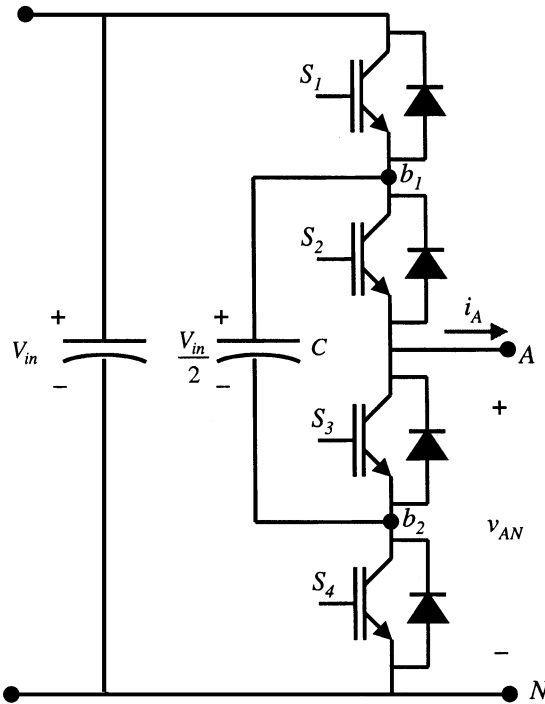


FIGURE 22-24 3-Level flying capacitor converter.

22.6 AC-DC CONVERSION: RECTIFICATION

AC-dc converters, or rectifiers, are used at the input of almost all line connected electronic equipment. Electronic devices that are powered directly from line and do not have regulation requirements use single- and 3-phase diode bridge rectifiers for converting line frequency ac to an uncontrolled dc voltage. For control over the output dc voltage thyristor-based rectifiers are used. Power factor corrected front end converters, discussed in Sec. 22.8, provide output voltage regulation as well as near unity power factor.

22.6.1 Single-Phase Diode Bridge Rectifier

Figure 22-25a shows the circuit of a single-phase diode bridge rectifier with a purely capacitive output filter. Due to its simplicity and low cost this circuit is preferred for low-power applications such as input stages of ac-dc adapters and computer power supplies. Diodes conduct in pairs to transfer energy from the input to the output when the input line voltage exceeds the output dc voltage in magnitude. Diodes D_1 and D_4 conduct when $v_s > v_o$, while D_2 and D_3 conduct when $-v_s > v_o$. The capacitor C_d gets charged by high current pulses during these small intervals near the peak of v_s , and discharges with the almost constant load current during the rest of the line cycle, as shown in Fig. 22-25b. The output dc voltage is approximately equal to the peak of the line voltage minus the forward voltage drop of two diodes. The capacitor value is chosen on the basis of the maximum load current and allowable output voltage ripple. The line current has significant harmonic content as shown in Fig. 22-25c. Source inductance of the line, common for regular utility supply, leads to lower peak input current, larger conduction times for the diodes, and reduced magnitude of the output voltage.

To quantify the line current distortion the following definitions are commonly used.

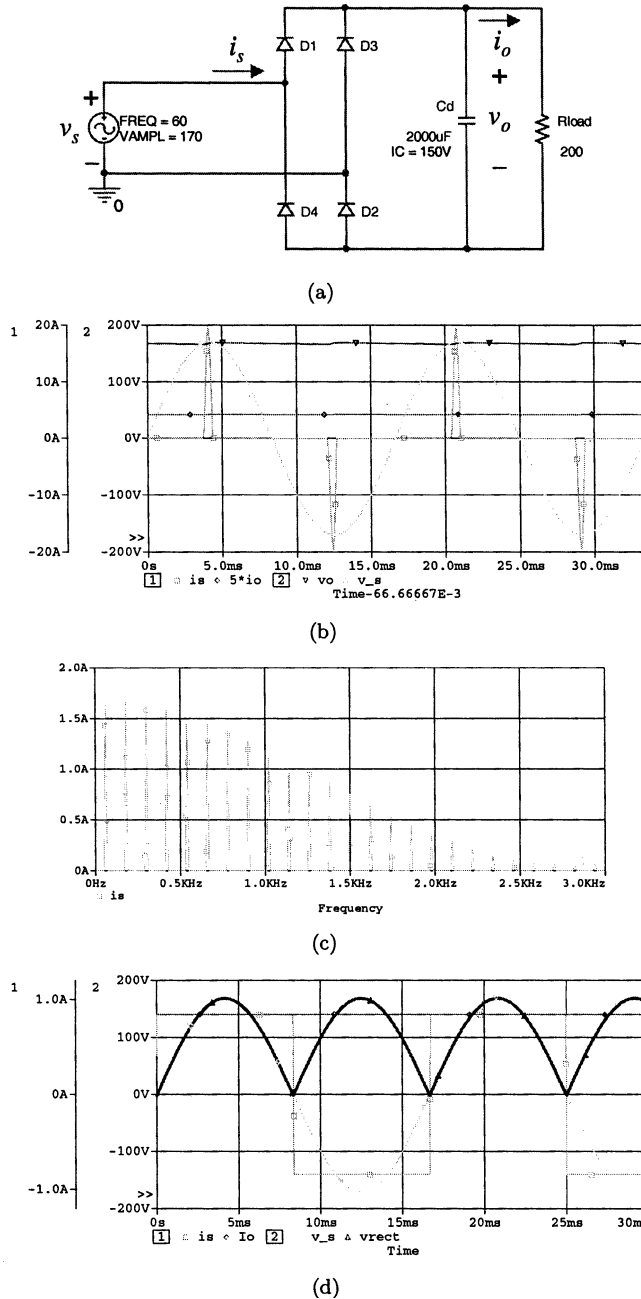


FIGURE 22-25 Single-phase diode bridge rectifier (a) circuit, (b) waveforms, (c) line current harmonics, (d) waveforms with inductive filter.

Total Harmonic Distortion. THD is the ratio of rms values of the distortion component to the fundamental component, expressed as a percentage

$$\text{THD} = \frac{I_{\text{dist}}}{I_1} \times 100 = \frac{\sqrt{I^2 - I_1^2}}{I_1} \times 100 \quad (22-49)$$

Real Power. This is the actual value of power consumed computed as an average over one line cycle

$$P_{\text{real}} = \frac{2\pi}{\omega} \int_0^{2\pi/\omega} v_s(t)i_s(t) dt = VI_1 \cos(\phi_1) \quad (22-50)$$

Apparent Power. It is the product of the rms values of the input voltage and current

$$P_{\text{app}} = V \cdot I \quad (22-51)$$

Power factor. Power factor (PF) is defined as the ratio of real power to apparent power.

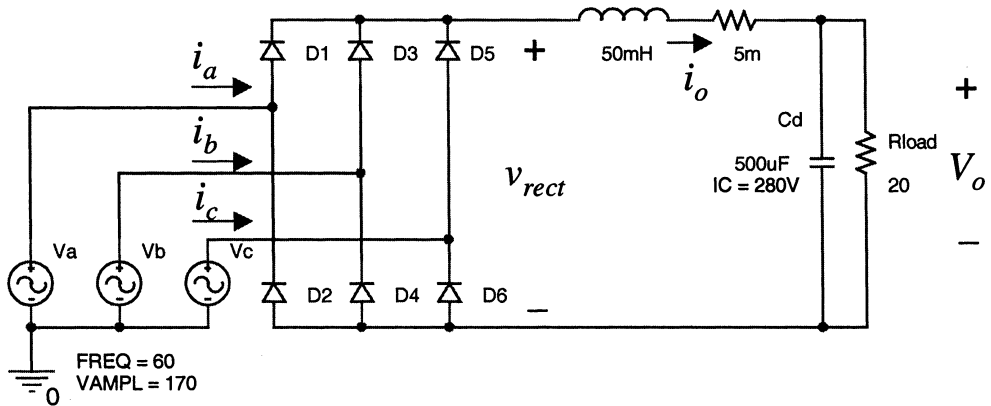
$$\text{PF} = \frac{P_{\text{real}}}{P_{\text{app}}} = \frac{VI_1 \cos(\phi_1)}{VI} = \frac{I_1}{I} \cdot \cos(\phi_1) \quad (22-52)$$

where V , I , and I_1 denote the rms value of the voltage, current, and fundamental component of the current, respectively, ϕ_1 is the phase angle of the fundamental component of the current with respect to the input voltage (assumed sinusoidal), and I_{dist} is the rms value of the distortion component of the input current. The term $\cos(\phi_1)$ is called the *displacement power factor*, while the term I_1/I is called the *distortion power factor*.

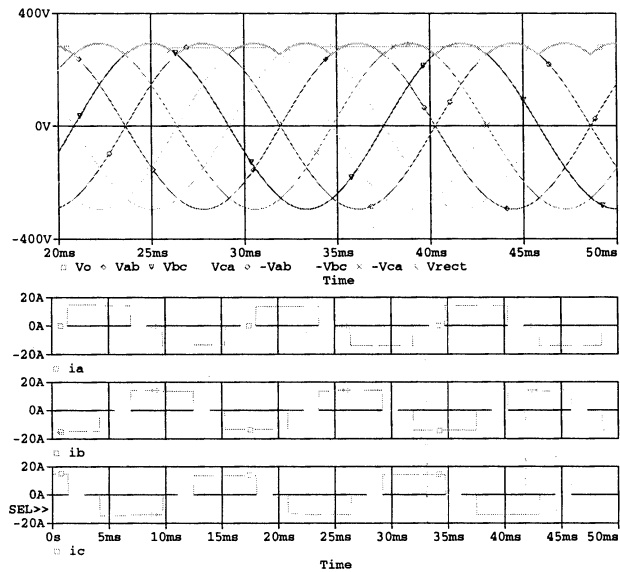
For the circuit values of Fig. 22-25a, the load current is 0.84 A, the peak line current $\hat{I}_s = 19.4$ A, rms line current $I_s = 3.8$ A, rms of the fundamental component $I_{s1} = 1.18$ A, $\text{THD} = 280\%$, and the $\text{PF} = (1.18/3.8) \cdot \cos(2.0^\circ) = 0.31$. The quality of the input current can be improved significantly if an inductive filter is used at the output of the rectifier. With a high enough inductance, the output current can be maintained nearly constant. This leads to a square wave shape for the input current as shown in Fig. 22-25d, which has a THD of 48% and a PF of 0.9. With the inductive filter, the output voltage has an average value equal to the average value of a rectified sinusoid, that is, $2\hat{V}_s/\pi$, where V_s is the peak value of the input phase voltage. Inductive output filter is preferable for medium power applications so that the input current has lower harmonic content.

22.6.2 Three-Phase Diode Bridge Rectifier

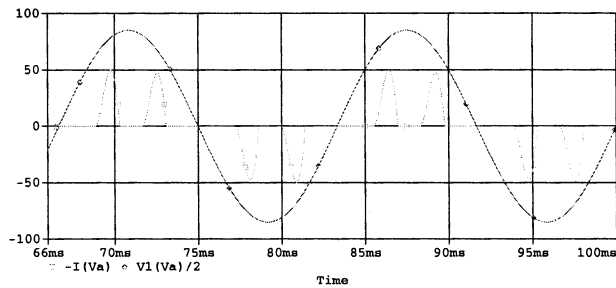
Figure 22-26a shows a 3-phase diode bridge rectifier with an inductive output filter. The operation is similar to the single-phase case. Diodes conduct in pairs—one from the upper three and one from the lower three. Cathodes of diodes D_1 , D_3 , and D_5 are connected together, so the diode with the highest voltage at its anode conducts. The converse holds for diodes D_2 , D_4 , and D_6 . The rectified voltage follows the envelope of the line voltages and their negatives: $v_{\text{rect}} = \max(|v_{ab}|, |v_{bc}|, |v_{ca}|)$. This rectifier is also called the 6-pulse rectifier because the voltage at the output of the diode bridge, v_{rect} , has six pulses in every line cycle. The average output voltage across the load is $V_o = \bar{v}_{\text{rect}} = (3/\pi)\hat{V}_{LL}$, $\hat{V}_{LL}$ being the peak value of the line-to-line voltage. The input line currents can be derived considering which diodes are conducting at a given time. They have quasi-square waveshapes as shown in Fig. 22-26b. The harmonic distortion is lower than in the single-phase case with inductive filter: $\text{THD} = 31\%$ and $\text{PF} = 0.955$. If the output filter is purely capacitive, the output voltage is equal to $\hat{V}_{LL}$, while the input currents are significantly distorted (Fig. 22-26c) and have harmonics at $(6m \pm 1)f$, where f is the line frequency and m is an integer. As in the single-phase rectifier with capacitive output filter, THD of the input current depends significantly on the source impedance.



(a)



(b)



(c)

FIGURE 22-26 3-Phase diode bridge rectifier (a) circuit, (b) waveforms with inductive filter, (c) line current waveform with capacitive filter.

The quality of input current and PF generally improve when going from single-phase to three phase, and can be further improved with higher number of phases if voltages with appropriate phase difference are generated from the utility supplied three phases. The output filter requirements also reduce as the number of phases is increased. With six voltage sources phase shifted by 30° , 12 diodes can be utilized to generate a 12-pulse rectifier. Isolated voltage sources phase shifted by 30° can be obtained using a wye-delta connected 3-phase transformer. Other phase shifts are generated by vectorial combination of appropriately scaled and isolated voltages that are obtained from the input 3-phase voltages using line frequency transformers. Rectifiers with pulse numbers 12, 18, and 24 are common for medium- and high-power applications that require good PF and low THD but do not have stringent constraints on size and weight.

22.6.3 Controlled Thyristor Rectifiers

Diode bridge rectifiers do not have any regulation capability and the output dc voltage varies with changes in line and load. This drawback is overcome by controlled thyristor rectifiers. Thyristor rectifiers are primarily used in medium- to high-power applications where regulation of the output dc voltage is required but line current quality and PF are not important (or can be corrected externally). Increasing concerns for power quality have resulted in reduced applications for these converters. High-power dc motor drives, especially those used in traction, battery chargers, and high-voltage dc (HVDC) transmission are the most common uses for these converters.

To understand the operation of thyristor rectifiers it is first necessary to know the basic terminal characteristics of thyristors. Thyristors, also called silicon-controlled rectifiers (SCRs), are high-power semiconductor devices that can block voltage of either polarity and conduct current in one direction (from anode to cathode). They can be switched on by applying a current pulse to their gate terminal when forward biased (positive voltage from anode to cathode), and can be switched off only by reducing the device current to zero.

Single-Phase Thyristor Rectifier. Figure 22-27a shows a single-phase *fully controlled* thyristor rectifier. The output has to be inductive for proper operation. For analysis presented here, it is assumed that I_o is constant and that there is no source impedance. During the positive half of the line cycle ($v_s > 0$), T_1 and T_4 are switched on after a delay angle α from the zero crossing of v_s . The angle α is commonly called the *firing angle*. With T_1 and T_4 on, $i_s = I_o$ and $v_{rect} = v_s$. When v_s reverses in

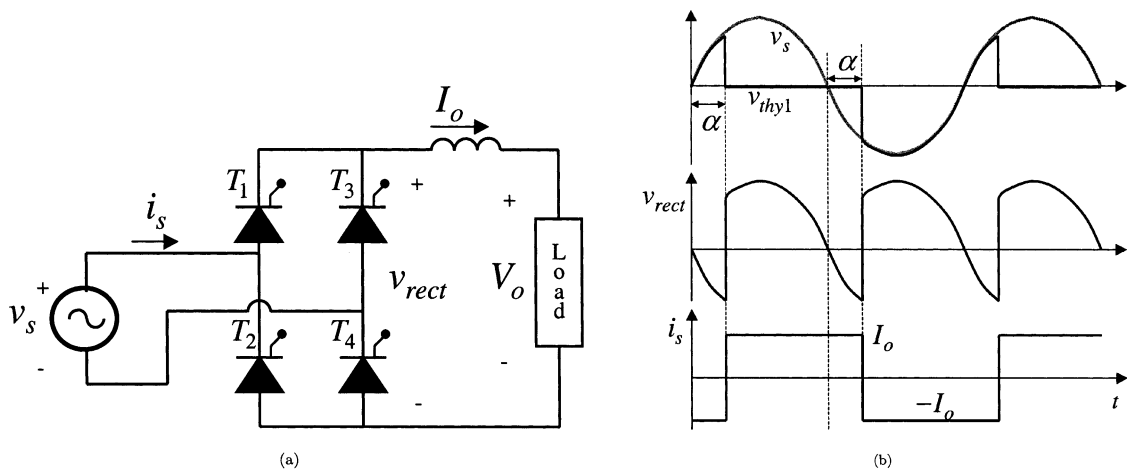


FIGURE 22-27 Single-phase thyristor rectifier: (a) circuit, (b) waveforms.

polarity thyristors T_1 and T_4 keep conducting since the current through them has not been reduced to zero. During the negative half cycle, T_2 and T_3 are switched on after angle α from the zero crossing of v_s . At this point current is transferred from (T_1, T_4) to (T_2, T_3) . In reality there is some finite source inductance, due to which the current transfer takes some time [4]. Once T_2 and T_3 start conducting $i_s = -I_o$, and there is a reverse voltage across T_1 and T_4 that keeps them in the off state. Steady state operating waveforms are shown in Fig. 22-27b. The average dc voltage across the load is given by

$$V_o = \bar{v}_{rect} = \frac{1}{\pi} \int_{\alpha}^{\alpha+\pi} \hat{V}_{ph} \sin(\omega t) d(\omega t) = \frac{2 \cos(\alpha)}{\pi} \cdot \hat{V}_{ph} \quad (22-53)$$

V_o can be controlled by varying $\alpha \in [0^\circ, 180^\circ]$. It is maximum for $\alpha = 0^\circ$, where the thyristor rectifier behaves exactly like a diode bridge rectifier, and zero for $\alpha = 90^\circ$. For $\alpha > 90^\circ$, $V_o < 0$, and power is transferred from the dc side to the ac side. Where bidirectional power flow is not required, thyristors T_2 and T_4 are replaced by diodes resulting in a *half-controlled* rectifier [43]. Total harmonic distortion of the input current is same as that in the diode bridge rectifier, but the fundamental component of the input current lags the input voltage by angle α , leading to a displacement PF of $\cos(\alpha)$. Thus, regulation of output voltage is achieved at the expense of lower PF.

Three-Phase Thyristor Rectifier. The 3-phase thyristor rectifier is shown in Fig. 22-28a. Similar to the single-phase case, each thyristor is switched with a delay of angle α after the anode to cathode voltage across it becomes positive. Each thyristor conducts for 120° , so the input line currents are quasi-square waves with magnitude equal to I_o , as shown in Fig. 22-28b. The average output voltage is $V_o = (3 \cos(\alpha)/\pi) \hat{V}_{LL}$. For $\alpha > 90^\circ$, power flows from the dc side to the ac side, and the converter acts like an inverter. For unidirectional power flow the three lower thyristors can be replaced with diodes to give a half-controlled version. As with diode bridge rectifiers, thyristor bridges can also be used to obtain 12 (or higher) pulse rectification resulting in lower THD for the input current and reduced size for the output filter. Due to the bidirectional power flow capability of this converter and very high voltage and current rating of thyristors, series connected thyristor rectifiers are utilized in HVDC transmission systems [44].

22.7 AC TO AC CONVERSION

In applications where a controllable 3-phase ac voltage has to be synthesized, the most common strategy is to first rectify line frequency ac to obtain a dc voltage, and then use a 3-phase inverter. The dc link requires a substantial electrolytic capacitor, which filters the dc voltage and also provides energy storage for short duration line voltage sags and interruptions. Capacitors add significant size and cost, and electrolytic capacitors also have the problem of lower reliability. To reduce the number of stages from two to one, and to eliminate the electrolytic capacitor, there has been a significant research effort in direct ac to ac conversion.

Thyristor-based *cycloconverters* [43] have been used extensively for direct ac to ac conversion. In these converters, a low-frequency ac waveform is synthesized by a piecewise combination of the available input ac voltages. These converters have been used for high-power variable frequency ac drives. However, they have limited control over the magnitude, frequency, and quality of the output voltage, and quality of the input line current.

Recently, matrix converters utilizing controllable bidirectional switches and PWM have been developed. As the name suggests, a matrix converter consists of a matrix of switches connecting each input phase to each output phase as shown in Fig. 22-29a. All the switches, denoted by square boxes in the figure, need to have bidirectional voltage blocking and current conduction capabilities. So far, a single semiconductor switch with these capabilities has not been invented. Thus, the switch has to be realized using a combination of existing power devices. One implementation is shown in Fig. 22-29b.

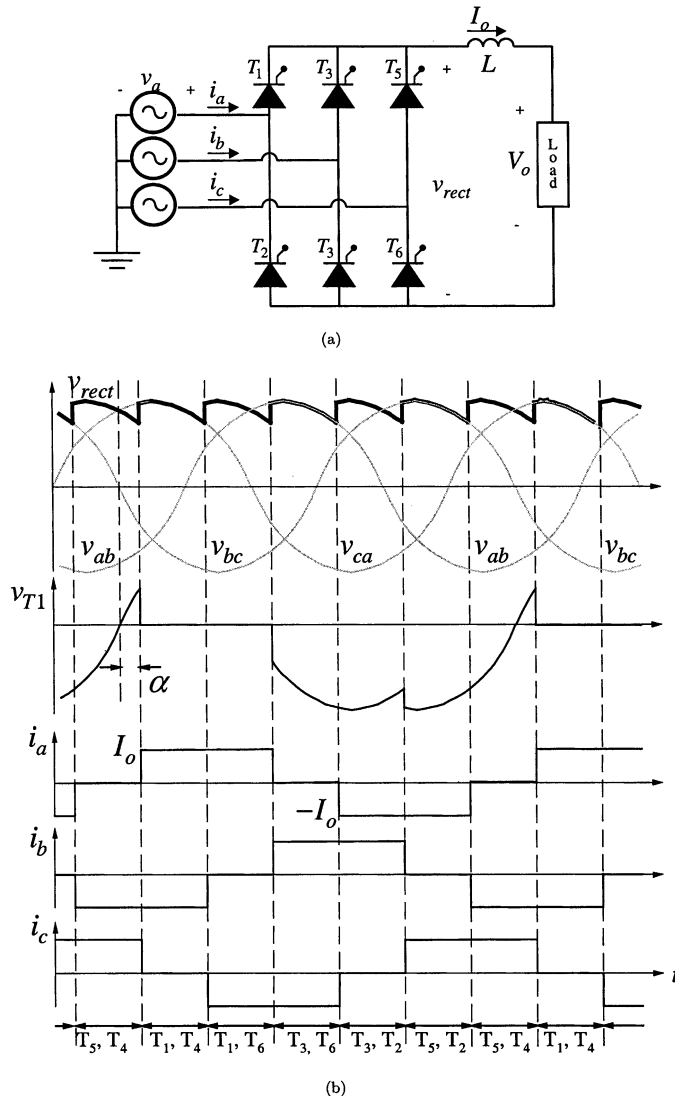


FIGURE 22-28 3-Phase thyristor rectifier: (a) circuit, (b) waveforms.

At any time instant each of the output phases is connected to one of the input phases, and more than one output phase may be connected to the same input phase. Selecting appropriate switches and using PWM, output voltages with continuously variable amplitude and frequency can be synthesized. The synthesis is most easily understood by means of space vectors [27, 45]. The total number of meaningful switching combinations are 27. Out of these, 6 lead to output voltage space vectors rotating at the input line frequency, 18 lead to stationary output voltage space vectors, while 3 lead to zero vectors. As in the dc to 3-phase ac case, on an average basis, a desired output voltage vector can be synthesized by using a combination of the stationary nonzero and zero space vectors. There is sufficient flexibility to ensure that the input power factor is unity. Considerable research has also

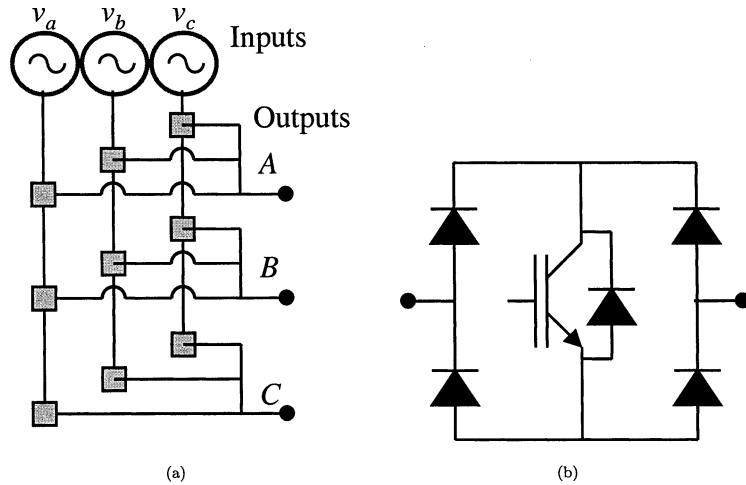


FIGURE 22-29 Matrix converters: (a) 3-phase matrix converter, (b) bidirectional switch.

been done to ensure proper operation of matrix converters under unbalanced line conditions. However, as shown in Ref. [46] the output voltage of a matrix converter has a theoretical limitation of $V_{LL,op} = (\sqrt{3}/2)V_{LL,in}$, which is considerably lower than that obtainable with an ac-dc-ac configuration ($V_{LL,in}$). In addition, clamping circuits and small input and output filters are required for proper operation.

So far, matrix converters have not been commercially successful. This is chiefly due to the number and cost of bidirectional switches, limitation on the maximum amplitude of the output voltages, and lack of energy storage, which is becoming increasingly important to provide *ride-through* capability during short duration line failures.

22.8 PROBLEMS CAUSED BY POWER ELECTRONIC CONVERTERS AND SOLUTIONS

The two main problems caused by power electronic converters are non-sinusoidal currents injected into or drawn from the utility and conducted and radiated electromagnetic emissions potentially leading to electromagnetic interference (EMI). In addition, power electronic converters have a negative incremental impedance, that is, as the input voltage reduces they draw higher current in order to supply a constant load power. Thus, as the total load supplied by power electronics increases, their effect on power system stability will also increase.

22.8.1 Harmonics and Power Factor Correction

As illustrated in Sec. 22.6, diode bridge rectifiers can inject significant current harmonics, and also result in a lagging power factor for the current. Both harmonics and lagging power factor lead to increased line losses. The harmonic currents in conjunction with the source impedance to the point of common coupling (pcc, where other loads are connected), lead to a distortion in the pcc voltage. Triplen harmonics from single-phase rectifiers contribute to zero-sequence currents when connected in a 3-phase system with a common neutral. If there are several single-phase diode bridge rectifiers connected across different phases and the neutral in a 3-phase system, it can lead to a very significant current flowing in the neutral wire, which is rated to carry only a small current due to load

imbalance. Grid connected inverters can also inject currents at the switching frequency unless the switching frequency is sufficiently high or appropriate filters are added.

Due to these concerns, standards have been formulated to limit the amount of distortion in current drawn from and injected into the utility supply. The IEEE 519 [47] specifies the maximum value for individual current harmonics and the overall THD for different applications and power levels. In addition, it also specifies a limit on the distortion in the pcc voltage. However, so far, the IEEE 519 is only a recommendation and not enforced by law. The IEC 1000-3 standard [48] specifies similar limits, and has been made into the EN61000-3-2 European norm. Thus, all electronic equipment sold in Europe has to comply with it. As a result, single-phase and 3-phase unity power factor (UPF) rectification techniques and power factor correction methods have been developed. Unity power factor rectification implies that the rectification technique ensures unity power factor operation. Power factor correction implies that the equipment itself does not draw input current at unity power factor; however, an additional circuit, such as an active, passive, or hybrid filter, is added to ensure that the line current is sinusoidal and in phase with the input voltage. The simplest single- and 3-phase UPF techniques are discussed below. A more comprehensive treatment can be found in Refs. [5, 49].

Single-Phase Boost UPF. The single-phase boost UPF, shown in Fig. 22-30a, is the most popular circuit for shaping the input current to be sinusoidal, maintaining it in phase with the input voltage, and regulating the output dc voltage [4, 5].

Basic Operation. The circuit consists of a full-bridge diode rectifier followed by a boost converter. The input voltage to the boost converter, v_{rect} , varies based on the diode conduction; ideally it is a rectified sinusoidal waveform. The output voltage v_o has to be greater than $\max(v_{rect}(t))$ for proper operation; to work with both 120 and 220 V ac inputs (also called universal input voltage range) the output is usually chosen in the range of 380 to 400 V. The boost converter operates so that there is a quasi-steady condition at each point of the input sine wave. Thus, the duty ratio varies as

$$d(t) = 1 - \frac{v_{rect}(t)}{v_o} = 1 - \frac{\hat{V}_s |\sin(\omega t)|}{v_o} \quad (22-54)$$

The duty ratio is controlled to satisfy two objectives: regulation of output voltage and shaping the inductor current to follow the waveshape of $v_{rect}(t)$. This is done by using the cascaded control structure shown in Fig. 22-30b. The error compensator for the outer voltage loop, $G_v(s)$, generates a reference signal for the amplitude of the inductor current (or the rectified input current), $\hat{I}_{L,ref}$. The effect of input voltage magnitude is fed forward by dividing the output of $G_v(s)$ by a signal proportional to square of the rms input voltage. The amplitude reference $\hat{I}_{L,ref}$ is multiplied by the waveshape of the rectified voltage to obtain $i_{L,ref}(t)$, the reference signal for the inductor current. The inductor current $i_L(t)$ is controlled to $i_{L,ref}(t)$ by using average current mode control. Steady-state waveforms are shown in Fig. 22.30c.

Control Issues. The current control loop has a bandwidth of the order of 10 kHz, required to follow the rectified sine wave shape accurately enough so that input current has acceptable THD. After each zero crossing of the input voltage, the positive rate of change of inductor current required is very high, while the voltage to effect this change, $v_{rect}(t)$, is close to zero. Thus, there may be a significant error in the current during this time, and the resulting distortion in the input current is called *cusp distortion*. Cusp distortion contributes to harmonics and lagging power factor.

The output voltage necessarily has a 120 Hz ripple component due to variation of power in single-phase systems. The error in the voltage feedback loop also has this ripple component. If this component is compensated by the control loop, it leads to a 120 Hz component in the output of $G_v(s)$, which after multiplication with the rectified voltage waveshape leads to a third harmonic component in the input current. To avoid this, the bandwidth of the voltage control loop is intentionally kept very low—in the range of 5 to 20 Hz. The converter, therefore, has very poor dynamic response with respect to changes in load. The input voltage feed forward signal also has to be filtered heavily to attenuate the ripple component sufficiently. Thus, although the feed forward improves the dynamic performance with respect to line changes, it is still quite slow. Further, to keep the voltage control design simple, a pure integrator is not added in the voltage

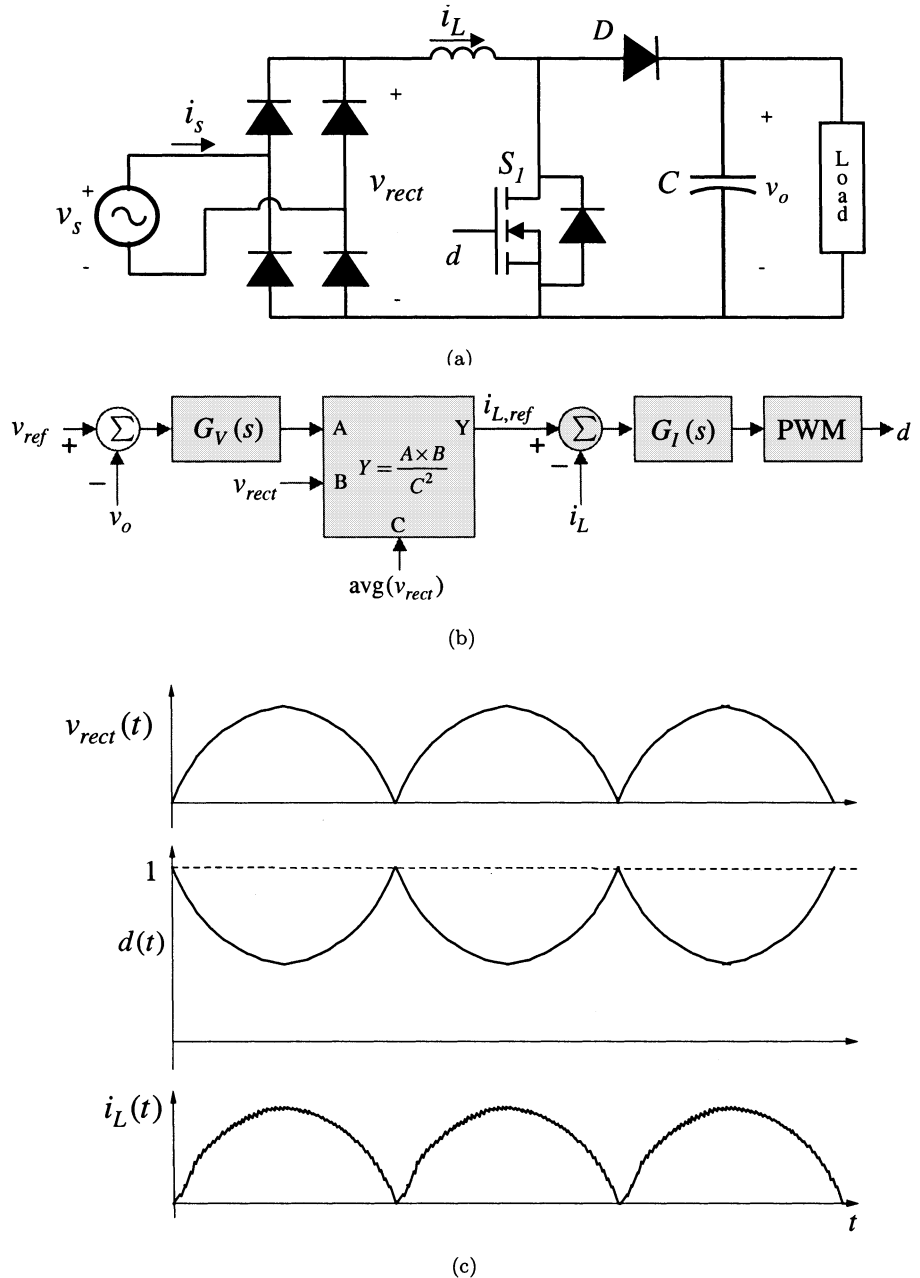


FIGURE 22-30 Single-phase boost UPF (a) circuit, (b) control block diagram, (c) steady-state waveforms.

control loop leading to a nonzero steady-state error. Typically, the output voltage drops by about 10 to 20 V from no load to full load.

The entire control, including the multiplier and divider circuit, can be implemented by means of a single analog integrated circuit (IC) such as the UC3854 [9], although digital implementations are also being explored. The availability of a single chip controller, easy to design power stage, and standardized control design equations (e.g., see the UC3854 application note on TI's website [9]) has made this topology very popular for single-phase applications. However, there are several other topologies that may be suitable for specific applications [5, 49].

The dynamic performance of the single-phase boost UPF is not adequate if the output is directly connected to a load. Usually, isolation of output from input line is a requirement, and the required output voltage may be significantly different from 400 V. Thus, an isolated dc-dc converter stage is required at the output of this UPF. This additional converter takes care of regulation, isolation, and scaling the voltage.

3-Phase UPF. In the 3-phase case, the diode bridge rectifier can be replaced by a 3-phase PWM inverter with external inductances in series with each phase. By appropriate control, this converter can transfer power from the ac side to the dc side. Input currents can be controlled in a manner similar to the single-phase case—waveshape determined by input voltages and amplitude determined by the voltage control loop. The converter operates in boost mode so the output dc voltage is regulated to a value higher than the peak of the input line-to-line voltage. In this case, dynamic response of the voltage control loop is limited by the nonminimum phase nature of the system. This 3-phase rectifier is very expensive compared to its diode bridge counterpart, and its cost is justified only if bidirectional power transfer capability is required. Thus, several alternate solutions for 3-phase power factor correction have been proposed [5, 49].

22.8.2 Electromagnetic Interference

Electromagnetic interference refers to maloperation of electronic equipment due to electromagnetic emissions which originate in the “surroundings”. Here, surroundings refer to other electronic equipment that are electrically connected or spatially close. Electromagnetic emissions emanate from transitions in currents and voltages, especially fast transients involving high magnitudes. Thus, power electronic converters can be major source of emissions leading to EMI. Within a converter the main sources are as follows:

- Current transfer between rectifier diodes.
- Switching of semiconductor devices: capacitive coupling between the high voltage switching terminals of the device and heat sink (which is usually grounded) leads to high-frequency common mode and differential mode currents.
- High-frequency inductors and transformers.
- Long wires or printed circuit board (PCB) traces carrying high-frequency currents.

Emissions can be classified into two types: *conducted* and *radiated*. In the conducted case, high-frequency currents produced due to switching transients flow through the parasitics in the power converter, and are manifested as common mode and differential mode currents at the inputs and outputs of the converter. At the output it can cause EMI problems for sensitive loads, and at the input it creates problems for other equipment connected to the utility. To restrict conducted emissions, EMI filters, good circuit board design, and packaging are needed [7, 50]. *Common mode chokes*, which offer high impedance to common mode currents and very low impedance to differential mode currents, are used to limit common mode currents. Differential mode high-frequency currents are attenuated by using capacitors to earth ground; the capacitance values are limited by the maximum allowable ground current specified by safety regulations like UL478 and UL1283. Complete EMI filters that provide both common and differential mode attenuation are commercially available as packaged units from several manufacturers.

Radiated emissions affect high impedance voltage sensitive circuits (e.g., MOS input circuits) that are in close proximity to the converter. To reduce radiated emissions, inductors and transformers are shielded (especially around air gaps), a careful PCB design that minimizes long paths for high frequency and switched currents is carried out, and long wires carrying high-frequency currents are twisted with their return lines. “Twisting” can also be approximately achieved for PCB traces. In addition, the entire converter may be shielded by a grounded metal enclosure.

Depending on the application power converters have to comply with an emissions standard. Emissions standards, formulated by the FCC [51], VDE [52], and the military, dictate the maximum amount of conducted and radiated emissions allowed for different kinds of electronic equipment.

22.9 APPLICATIONS OF POWER ELECTRONIC CONVERTERS

The foregoing sections described the principles and control of different power converters. This section describes the requirements of specific applications, and how power electronic converters are utilized for these. The next section is devoted to utility applications of power electronics.

22.9.1 DC Power Supplies

DC power supplies are required for powering different components in electronic equipment. Their specific features and level of sophistication depend on the application. For example, in a *bias supply*, required for various analog components in a power converter, the main requirements are isolation and a (relatively) large tolerance in the output voltage level. The input to the bias supply may be derived from a regulated dc bus, so line regulation is not required. A simple push-pull converter without an output inductor can be used for this purpose. For use inside a sensitive instrument, the power supply requirements would include stringent regulation with respect to line and load, good dynamic response, compliance with EMI and input power quality standards, and adequate energy storage for normal operation during short duration line failures. Features commonly required in dc power supplies are as follows.

Output Voltage Regulation with Respect to Line and Load. For high-current low-voltage power supplies there may be an additional requirement of regulating the voltage at the load terminals. This requirement, called remote sensing, accounts for voltage drops in the connecting wires. Dynamic requirements of response time and overshoot/undershoot in the output to step change in load current are also specified.

Output Current Limit. This may be a fixed value, or have a *foldback* characteristic [6]. With the latter, once the output current exceeds a certain value, the current limit is varied as a function of output voltage—decreasing as the output voltage decreases.

Isolation. Electrical isolation between output and input.

Soft Start. This is required to limit the inrush current for initial charging of capacitors.

Holdup Time. For line powered applications, there is usually a requirement for holdup time, time for which the power supply should operate normally in the absence of the input voltage. Usually energy storage is provided by adding capacitors; sometimes auxiliary capacitors charged to a higher voltage to store more energy are used.

Sleep Mode. Typically power supplies have a low efficiency at light load. Thus, in battery-powered and portable applications light load condition is detected and the power supply operation is changed to reduce its losses, for example, by reducing the switching frequency.

Power Factor Corrected Front End. For single-phase applications, it also helps in operation with universal input voltage range (100–240V, 50/60 Hz).

EMI Compliance. With a specified standard.

Environmental compliance. For temperature, humidity, and altitude of operation. Some features can be implemented with standard PWM control ICs leading to reduced number of components and design simplification. Commercial availability of low power dc-dc converters modules with standard input and output voltages has increased significantly in recent years. Thus, a custom power converter design may not be required, unless there are special requirements as in space, defense, and some instrumentation applications.

A specific application of dc power supplies is in digital systems. All digital systems need a power electronic converter to provide the requisite supply voltages. With the digital supply voltages going down and the clock frequencies going up, the requirements are increasingly toward low voltages and high currents. For a high-speed microprocessor, the required values are around 2 V and a few hundred amperes. The dynamic requirements are also very demanding since the microprocessor load can go from almost zero current to full load current in a few microseconds. Furthermore, digital components operating at different supply voltage levels may be used in a complete digital system. For these applications, the intermediate bus architecture is used. First, a 12 V (or a lower voltage) bus is derived from the input supply using an intermediate bus converter (IBC); the IBC may also provide isolation from the input. This low-voltage bus is then input to several point-of-load (POL) converters, which are located close to the loads and provide the required steady state and dynamic voltage regulation. Point-of-load converters do not provide isolation between input and output. Intermediate bus converter and POLs with standard input and output voltages are commercially available as modules from several manufacturers (e.g., see Refs. [17, 53]).

22.9.2 Electric Drives

Introduction. Traditionally electric motors have been powered by direct connection to ac line, or to dc voltage obtained from a rectifier. However, this usually results in inefficient operation due to lack of control. Electric motors powered by appropriately controlled power electronic converters lead to significant increase in the overall system efficiency due to the advantages of variable speed operation [54]. Electric motor load has more than half the share of electric power consumption in the United States, of which about half is in industrial applications. Thus, any increase in system efficiency due to electronic controlled motor drives can lead to large savings for the company and overall electricity consumption. Moreover, performance advantages of fast dynamic response and very accurate control over speed and position are obtained.

Squirrel cage induction machines (IMs) are the most widely used electric motors. This is due to the advantages of simple, robust, and low-cost construction, ease of powering, and lower maintenance requirement due to the absence of brushes. Thus, control of induction machines for better performance and improved efficiency has been researched extensively. The invention of vector control [33], and direct torque control [34, 35] have made the dynamic performance of induction machines similar to that of a dc machine. The popularity of vector control was also aided by development of PWM inverters and space vector modulation described in Sec. 22.5.3. Developments in permanent magnets and power electronics, and increasing concerns for size, weight, and efficiency have led to significant interest in permanent magnet synchronous machines (PMSM) and brush-less dc (BLDC) machines. Permanent magnet synchronous machines and BLDC motors have the highest operating efficiency and the highest power density of all the motors [55]. Further, due to development of power electronics, the switched reluctance machine (SRM), the first electromechanical machine invented in the early 1800s, has also been commercialized [56, 57]. This machine has a simple, robust, and inexpensive construction, and is powered from an inherently robust power converter topology. Switched reluctance machines are most popular in the low-to-medium power range and high-speed applications requiring medium performance. Squirrel cage induction, PMSM, Synchronous Reluctance,

and BLDC drives require a 3-phase inverter for powering the stator windings and no electrical excitation for the rotor. Switched reluctance machines have a special power converter that has independent sections for individual phases. For illustration, a vector controlled ac drive and an SRM drive are discussed here. Details of these drives may be found in Refs. [28, 30, 36, 37, 54–57].

Vector Controlled AC Drive. Figure 22-31 shows the basic block diagram of a vector-controlled PMSM drive. The structure is identical for the squirrel cage induction motor drive. The input stage is a single- or 3-phase diode bridge rectifier, the output of which is filtered to obtain a dc. The input current during initial charging of the dc bus capacitor is limited using either (1) parallel combination of a controllable switch and a current limiting resistor connected in series with the main power path, or (2) a half-controlled thyristor rectifier. The dc voltage is converted to variable frequency variable magnitude 3-phase ac by means of a 3-phase PWM inverter. During *dynamic braking*, energy stored in the inertia of the machine and load is transferred back to the electrical circuit. If the rectifier has bidirectional power transfer capability, this energy can be transferred to the source. Otherwise, the energy charges the dc bus capacitance. To limit the dc bus voltage, a brake resistor is connected in parallel with the dc bus by means of a controlled switch.

The switching frequency of the inverter in ac drives is usually between 20 and 30 kHz, since the machine inductance is usually sufficient to filter the output current. Thus, IGBTs are commonly used for this application. MOSFETs are used if the dc voltage is below 600 V and a higher switching frequency is required in order to reduce the output current ripple. Intelligent power modules (IPMs) containing all the power semiconductor components along with gate drive and protection circuits in a single package are now available from several manufacturers (e.g., Refs. [58–62]). Use of IPMs results in considerable design simplification, size reduction, and smaller design cycles.

The drive control is carried out using a microcontroller or digital signal processor (DSP), with the use of DSPs becoming more popular due to increased computational requirements. In an industrial application, the DSP controls the machine in response to command signals (for speed or position) obtained over a communication interface bus. AC machines are usually controlled using either vector control or direct torque control. From the power electronics point of view, this requires control of the output currents with a high control bandwidth (of the order of 1 kHz). This is done in a cascade manner: the current control algorithm generates reference voltage vectors, which are then synthesized using space vector PWM. Both current control and PWM are implemented on the DSP. The final output of the DSP are switching signals for each of the three inverter legs and the brake switch. These signals are optically isolated and input to the gate drive circuitry for each of the

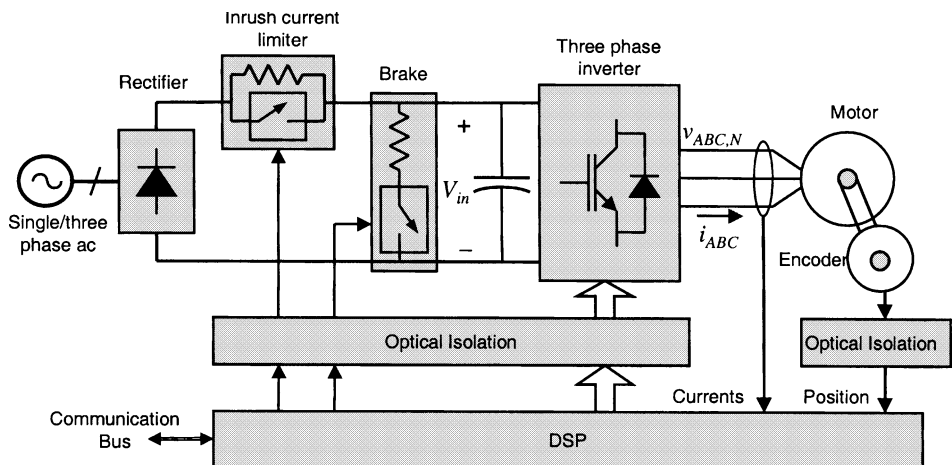


FIGURE 22-31 Vector controlled 3-phase ac drive.

controlled switches. Output currents of the inverter are sensed using either hall effect sensors or sense resistors. The analog current signal is converted to digital form using the internal analog to digital converters (ADCs) of the DSP. The dc bus voltage may also be sensed for feedforward of input line variations and dynamic braking. An optical encoder is used if direct rotor position sensing is required. Since the encoder outputs are already in digital form they can be easily interfaced. However, due to close proximity of the encoder and the machine windings, which are subject to pulsed voltages generated by the inverter, these signals have very high common mode noise. To solve this problem, differential mode line drivers and receivers, and optical isolation of encoder signals are commonly employed.

SRM Drive. The basic structure of an SRM drive is similar to an ac motor drive. However, the power converter is quite different owing to the machine characteristics. Figure 22-32a shows the cross-section of a 4-phase SRM. The SRM has saliency on the stator as well as the rotor. Each phase consists of concentrated coils wound on diametrically opposite stator poles. If phase a carries current i_a then the closest set of rotor poles are attracted to the stator phase a poles. Once the rotor poles are aligned with the stator poles, there is no torque on the rotor and i_a has to be reduced to zero. By energizing and deenergizing phases in sequence a continuous rotation can be achieved.

To build up and reduce winding current, the converter needs to have bidirectional voltage capability. However, since the direction of torque is independent of the current polarity, only unidirectional current capability is required. At low speeds, the back emf of the machine is very low so the winding current has to be controlled usually by hysteretic current control. This is usually done by letting the current freewheel with a zero voltage across the winding terminals, commonly called application of a zero voltage loop (ZVL). To determine which phase needs to be energized or deenergized rotor position information is required, this may be sensed or estimated, and discrete position information is sometimes sufficient [57]. Usually the total torque produced by the machine has significant ripple, and unless instantaneous torque control is a requirement, average torque control is implemented. The average torque depends on the turn-on and turn-off angles (rotor positions with respect to the phase winding where that phase is energized and deenergized, respectively), and the current magnitude if current control is used [56]. The control is usually implemented based on results of numerical simulation and then tuned experimentally.

Ray and Davis published one of the earliest paper on power converters for SRM [63], and suggested the asymmetric bridge converter, which is also known as the classical converter. The circuit topology, shown in Fig. 22-32b, is similar to a two-switch forward converter [4] and fulfils all the

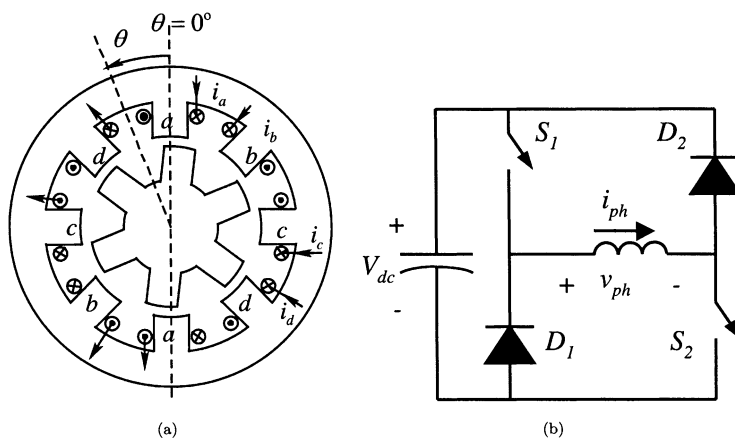


FIGURE 22-32 SRM Drives: (a) Cross-section of a 4-phase 8/6 SRM, (b) asymmetric bridge converter.

basic requirements. Only one phase is shown, the circuit being identical for the other phases with a common connection to the input dc bus. The basic operating modes of the converter are as follows:

Energization	(S_1, S_2) on, (D_1, D_2) off	$v_{ph} = V_{dc}$
Deenergization	(D_1, D_2) on, (S_1, S_2) off	$v_{ph} = -V_{dc}$
ZVL	(S_1, D_2) on, (S_2, D_1) off, or (S_2, D_1) on, (S_1, D_2) off	$v_{ph} = 0$

where all switches and diodes have been assumed ideal and the winding resistance has been neglected. This converter is extremely reliable and fault tolerant. The problem of “shoot-through” due to spurious turn-on of the two controlled switches in the same leg is avoided, since the phase winding is in series with the controlled switches. Shoot-through can be a major problem in full-bridge dc-dc converters and 3-phase inverters. Although this converter topology has become very popular and a packaged module similar to 3-phase ac inverters is now available, there are several other converter topologies, which may be suitable for specific applications. A detailed comparison of SRM converter topologies can be found in Refs. [56, 64–66].

DC motor drives use a dc-dc converter—buck, buck-boost, or full bridge depending on the requirements—and armature current control for controlling the instantaneous torque. The field current may also be controlled to reduce back emf at high speeds. Their popularity has been decreasing due to maintenance requirements and speed limitations of brushes. As the cost of power electronics reduces, better and low-cost permanent magnets are developed, better control and estimation techniques are implemented on lower cost DSPs, and energy efficiency gains priority, it is expected that PMSM and BLDC drives will increase their share in motor drives. Switched reluctance machines are also expected to gain ground in high speed and harsh environment applications where energy efficiency is not very important.

22.9.3 Battery Charging

Battery charging is a very large part of the ac-dc and dc-dc converter applications. The end use for batteries is in telecommunications, electric and hybrid electric vehicles, portable electronics, and energy storage for improving power system stability. For a lead acid battery, there are three standard operating modes for the battery charger: (1) constant current (or bulk) charging during low state of charge, (2) constant voltage charging after about 80% state of charge, and (3) “float” or trickle charge after the battery reaches its open circuit voltage. Thus, the requirement on the power converter is to operate at its maximum current rating over a significant output voltage range with high efficiency. In addition, isolation from input line and power factor correction are usually required.

In recent years, lithium-ion (Li-ion) batteries have gained widespread usage for portable applications like laptop computers, PDAs, and cell phones. They have also been proposed for automotive applications, but so far nickel metal hydride (NiMH) batteries are used in commercially available hybrid electric vehicles (HEV). The advantages of Li-ion batteries are high energy density, higher cell voltage of about 4.2 V, and low self-discharge rate [67]. The disadvantages are high sensitivity to electrical stress and limited temperature range. The cell voltage has to be monitored for overvoltage (which is very close to the nominal open circuit voltage) to avoid catastrophic failure, and undervoltage to maintain battery life. Linear chargers are used for single-cell batteries requiring low charging current, while switch mode converters are used for high-voltage and high-current charging. The charging technique is similar to lead acid—constant current followed by constant voltage. However, pulse charging (alongwith periodic discharge and periodic relaxation pulses) have also been proposed for improving battery life. A further complication with Li-ion is that the dependence of state of charge on open circuit cell voltage is only on a small voltage window (3.2–3.8 V). Thus, in the absence of a detailed cell model, a coulomb counting technique, which measures the charging and discharging current and estimates the self-discharge rate, is normally used to determine the battery state.

22.9.4 Fluorescent Lamps and Solid State Lighting

About 30% of the total electricity generated in the United States is consumed for lighting [68]. Thus, there is significant interest in increasing energy efficiency of lighting mechanisms. Incandescent lighting, although the most inefficient, is still the cheapest in terms of \$/lumens. However, compact fluorescent lamps (CFLs) have gained ground due to their energy efficiency and are now commercially available as replacements for incandescent bulbs. Besides lighting, fluorescent lamps that emit ultraviolet light are used in several industrial applications like sterilization and curing (drying of coatings).

Compact fluorescent lamps and industrial lamps are usually powered using a switched mode converter called an electronic ballast. In one implementation, the converter consists of a full- or half-bridge inverter with its output connected to an L-C-C (all in series) tank circuit. The dc voltage input to the inverter may be obtained using either a diode bridge rectifier or a power factor corrected front-end. The lamp filaments are connected in series with the L-C-C tank, with the two filaments being in parallel with one of the capacitors. The other capacitor provides dc blocking and also tunes the tank operation to a desirable characteristic. When the lamp is off (gas inside not ionized), it acts as an open circuit, so the inverter is effectively connected to a series resonant L-C circuit. To ionize the gas inside the lamp, the filaments have to be heated sufficiently followed by application of a high voltage (~kV) across the lamp for a short time. Once the lamp is on (gas ionized) it acts a resistive load, and the inverter load is then a series parallel tank circuit—series L-C-C with the equivalent resistance of the lamp in parallel with one capacitor. The inverter outputs a high-frequency square wave, typically in the range of 20 to 100 kHz; lamp operation at these frequencies results in increased light output. The switching frequency is varied to control the ballast operation. Initially, the tank circuit and the lamp are excited at a frequency significantly higher than the resonance frequency. During this time, called the preheat time, the lamp filaments increase in temperature. After a suitable delay, the switching frequency is reduced toward resonance so that the voltage across the lamp increases until the gas inside the lamp ionizes and the lamp “ignites”. After ignition, the lamp acts like a resistive load, and the lamp output power is controlled indirectly by regulating the phase difference between the inverter current and voltage. The tank circuit also ensures resonant transitions of the converter switches [69] leading to reduced losses. Several ICs that implement full control of a fluorescent lamp including preheat, ignition, and dimming control are commercially available (e.g., see Ref. [62]).

High intensity discharge (HID) lamps have a large share of commercial lighting such as street lighting, sports facilities, etc. They are of three types—mercury vapor, sodium vapor, and metal halide. The ballast requirements of these lamps differ from those of fluorescent lamps. Operation of these lamps at higher frequencies does not improve light output. Further, they exhibit acoustic resonance in the 10 to 100 kHz range. Thus, so far, electronic ballasts for these lamps have not become very popular.

Power supplies for plasma cutting have requirements similar to fluorescent lamp ballasts. Initially, a high voltage (~10 kV) is applied to produce an arc between two high-voltage electrodes. This is followed by *arc transfer*, transfer of the conducting ionized gas between the electrodes to the space between the negative electrode and the metal being cut (workpiece). Once the arc is transferred the cutting procedure requires a current controlled supply (at lower voltage) connected across the negative electrode and the workpiece. Since there are three terminals, two separate power supplies can be used—one for generating the initial high voltage and the other for the actual cutting operation.

The U.S. Department of Energy has identified solid-state lighting as a means to increase lighting efficiencies in the near future [68]. Recent and expected breakthroughs in semiconductor and organic light emitting diodes (LEDs) are the main impetus behind this. Another reason promoting the use of LEDs is the European Union’s directive on removal of hazardous substances (RoHS) in electrical and electronic equipment (Directive 2002/95/EC). The directive aims at phasing out the use of mercury and lead (among other substances), which are chief components of compact fluorescent and HID lamps. LEDs do not use either of these materials. Although the RoHS directive, going into effect in July 2006, makes exceptions for fluorescent and some HID lamps used for lighting, it is

expected that these will be covered in future revisions of the directive. Light emitting diodes are already being used for traffic lights, exit signs, and as indicator lights in automobiles. They have the advantages of faster turn-on time, increased efficiency, vibration and shock resistance (required for automotive), and longer operating life. For solid-state lighting, light output of several series connected LEDs, called *strings*, is used. A switched mode converter with a dc current regulated output is used to power the LED string. The output current is regulated to provide dimming control and to avoid failure due to overcurrent [70]. When LEDs start replacing HID lamps, incandescent bulbs, and CFCs for commercial/residential lighting, they will open up a huge market for efficient and low-cost power electronic converters.

22.9.5 Automotive Applications

In recent years, electronically controlled load in automobiles has increased significantly. Further increase is expected due to more comfort features, and potential replacement of some mechanical systems by all electrical systems like active suspension and power steering. The present bus voltage in automotive systems (14 V), decided by the alternator charging voltage, implies very high currents for the expected high power consumption. To keep the current levels manageable, the 42 V power net has been proposed, and adopted by most auto companies and suppliers for future generation automobiles [1]. The choice of 42 V has been made on the basis of safety considerations, load dump over-voltage transients, and optimal utilization of silicon in power semiconductor devices. It is expected that there will be a dual voltage system, with both 42 and 14 V loads, although the exact configuration has not been standardized. Thus, a dc-dc converter, possibly bidirectional, will be required for interconnection between the two system voltages. Furthermore, converters with sophisticated controls will be required for features like active suspension and power steering [71].

Hybrid electric vehicles (HEVs) are now gaining commercial success due to their higher mileage, lower emissions, and tax breaks. The basic ideas in an HEV are:

- Operate engine at optimal efficiency, achieved at high speeds.
- During initial acceleration and at low speeds, power is supplied by a battery powered motor leading to reduced idling losses and emissions.
- For high acceleration, power is derived from both the engine and the battery.
- During high speeds, the battery is charged from a generator connected to the engine. The motor used for generating mechanical power can also be used as a generator.
- During braking, the generator feeds energy back to the battery.

Power electronic converters are required to control the motor (and generator for a separate machine), and for battery charging. In HEVs, the battery has to be maintained at less than full charge to receive regenerative energy during braking. Hybrid and completely electric vehicles with fuel cells as the power source and some means of energy storage are also being developed by several auto manufacturers. These require power converters to interface fuel cell, energy storage element, motor/generator, and other electronically controlled loads.

22.10 UTILITY APPLICATIONS OF POWER ELECTRONICS

22.10.1 Introduction

Power electronics has the potential to change the landscape of power generation, transmission, distribution, and end use. Growing energy demand, coupled with economic, environmental, and political restrictions on newer generation and transmission infrastructure, means that the existing

resources operate near their stability limits. As a result dynamic instability, inter-area oscillations, voltage instability cascading to major blackouts have become issues of real concerns today. Power electronic technology is widely considered to be one of the key components of the grid modernization.

Distributed generation is an important energy option for the twenty-first century and is a key element of restructuring of the electric utility. Distributed generation using renewable sources such as solar, wind and tidal energy, and hydrogen, and their interface through power electronic converters, represents one of the most promising paths to sustainable energy. Also, several modern loads such as the processing plants in semiconductor industry or data centers require clean and uninterrupted power. Power electronic-based power quality solutions are essential for these loads to mitigate problems such as voltage sags, harmonics, and flicker in line voltage.

Widespread use of power electronics in power systems is further fueled by dramatic advances in power semiconductor materials and devices, especially those based on silicon carbide (SiC) [72, 73], and advances in the fields of wide area power system monitoring and communication. This section briefly describes the major power electronics applications in power systems, namely, flexible ac transmission systems (FACTS), custom power, and interfacing distributed generation with electric grid. Figure 22-33 shows the interconnected power system network with some of the major power electronics highlighted.

22.10.2 Flexible AC Transmission Systems

Flexible ac transmission systems is a collective term for different types of power electronic devices/converters based systems that are capable of controlling power flow in high-voltage ac transmission systems [74, 75]. With advances in power semiconductor devices, PWM methods, and control theory, the use of FACTS devices has seen a significant increase. Flexible ac transmission systems devices are capable of the following major functions:

- Control power flow along desired transmission corridors, which is critical for a deregulated utility; they can also minimize loop flows.
- Increase transmission capacity without requiring new transmission infrastructure.
- Improve transient, dynamic, and voltage stability, and provide damping for inter area oscillations.

Different types of FACTS devices control different parameters of the transmission system like the effective line impedance, bus voltage magnitudes, or phase angles, to control power flow, and to

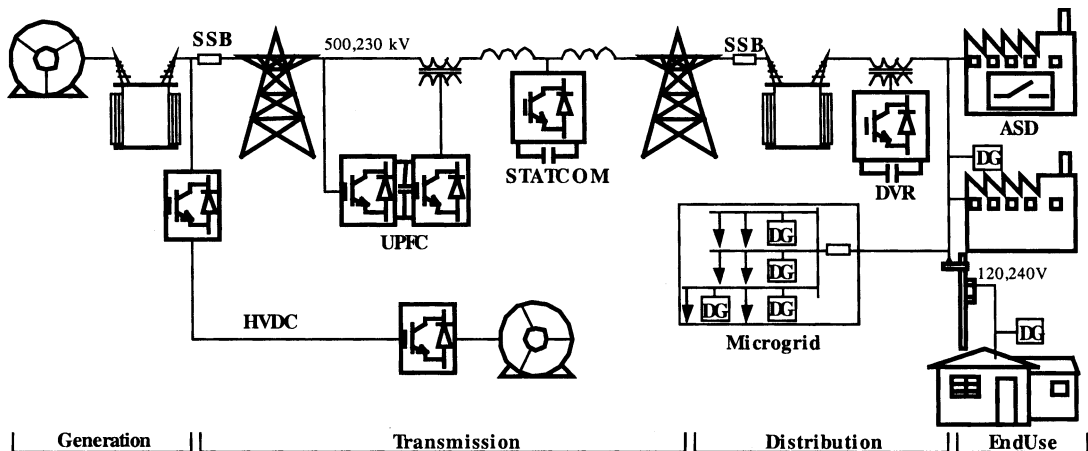


FIGURE 22-33 Power electronics applications in power systems.

increase stability margins. Consider the single line diagram of a 2-bus system shown in Fig. 22-34. The real power flow in the connecting transmission line is given by

$$P = \frac{V_1 V_2}{X} \sin(\delta) \quad (22-55)$$

where V_1 , V_2 are the magnitudes of sending and receiving end voltages, respectively, δ is the phase angle between the two voltages, and X is the series line impedance [76]. Flexible ac transmission systems devices control one or more of these three parameters to control power flow and improve stability.

Transient Stability. Flexible ac transmission systems devices have the ability to enhance both transient and dynamic stability of power system networks, thereby enabling increased power flow through existing transmission lines. Transient instability occurs when a major disturbance like fault, line outage, or loss of generation results in large rotor angle deviations leading to loss of synchronism. The rotor angle deviation is governed by the swing equation given in Eq. (22-56).

$$\frac{2H}{\omega_o} \frac{d^2\delta}{dt^2} = P_m - P_e \quad (22-56)$$

where H is the inertia constant in MJ/MVA, ω_o is the synchronous speed, while P_m and P_e are the mechanical power input and electrical power output, respectively. During a fault, the electrical energy drawn from the generator reduces significantly, while mechanical power input remains roughly constant, leading to increasing rotor angle. If the fault is not cleared before a critical time, transient instability occurs. The critical clearing time depends on the electrical power output during the fault and immediately after fault clearance. Since power flow can be controlled continuously using FACTS devices, power during and after a fault can be controlled to improve the stability margin of the system.

Transient instability is often studied using the equal area criterion as shown in Fig. 22-35 [76]. Initially the mechanical input power input is equal to the electrical power transmitted at an angle δ_1 . A fault at the generator makes the electrical power zero while the mechanical input power remains the same, leading to increase in rotor angle from δ_1 to δ_2 , at which point the fault is cleared. During this interval, the stored kinetic energy in the machine increases, and increase in kinetic energy is equal to the area A_1 in Fig. 22-35. After the fault is cleared the electrical power transmitted is higher (due to increased phase angle) than the mechanical power input. Hence, the machine begins to decelerate. However, the phase angle increases further due to the stored kinetic energy. The maximum angle is reached at δ_3 , when the decelerating energy represented by the area A_2 becomes equal to the accelerating area A_1 . If the phase angle extends beyond δ_{crit} , then the system is unstable since decelerating energy cannot balance the accelerating energy. The area A_{margin} between δ_3 and δ_{crit} , represents the transient stability margin of the system. Flexible ac transmission systems devices can improve the margin by dynamically changing the $P - \delta$ characteristics of the system.

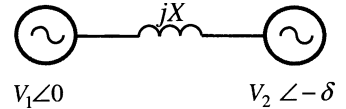


FIGURE 22-34 Single line diagram of a two-bus system.

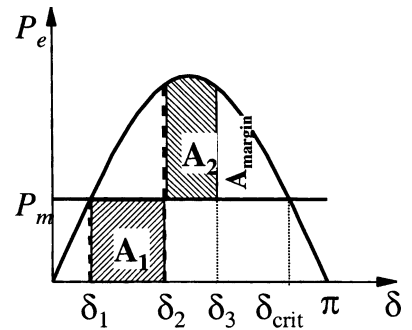


FIGURE 22-35 Equal area criteria for transient stability.

Thyristor Controlled Series Capacitor. The earlier FACTS devices were predominantly thyristor based, like the thyristor-controlled series capacitor (TCSC) and static VAR compensator (SVC) [75]. Thyristor controlled series capacitor is a series-connected FACTS device that controls the effective impedance

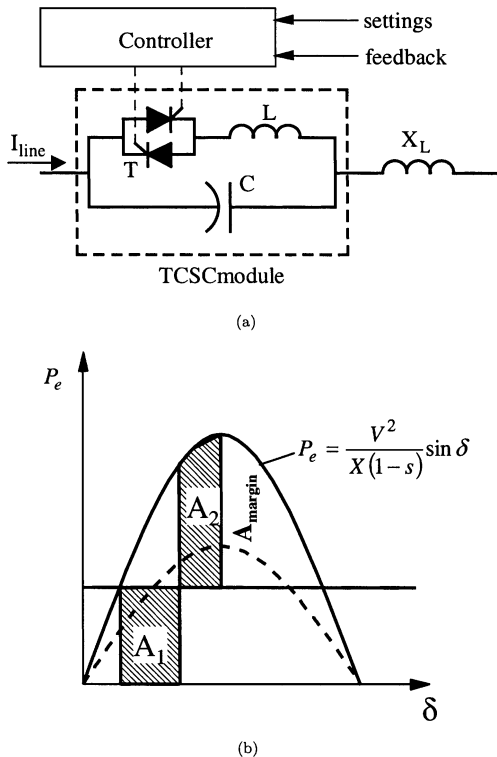


FIGURE 22-36 TCSC: (a) schematic (b) enhancement of transient stability.

of the transmission line. Figure 22-36a shows the basic schematic of a TCSC, which consists of a capacitor in parallel with a thyristor-controlled reactor (TCR). Thyristor controlled reactor is a series combination of an inductor and a pair of phase-controlled thyristors. By suitably controlling the firing angle of the thyristors, the reactance (inductive) of TCR, and therefore, the effective fundamental impedance of TCSC can be controlled continuously. The relationship between the firing angle α is measured from the zero crossing of the capacitor voltage. As an example, corresponding to an installed capacitive impedance of 0.5 pu (per unit) and inductive reactance of 0.1667 pu, the effective impedance of TCSC can be controlled from about 4 pu capacitive to 2 pu inductive. The effect of TCSC control on the transient stability margin is illustrated in Fig. 22-36b. In the figure, s indicates the degree of compensation. One of the major advantages of TCSC, when compared with uncontrolled series compensation, is the ability to mitigate sub synchronous resonance (SSR) [75].

Voltage Source Converter Based FACTS.

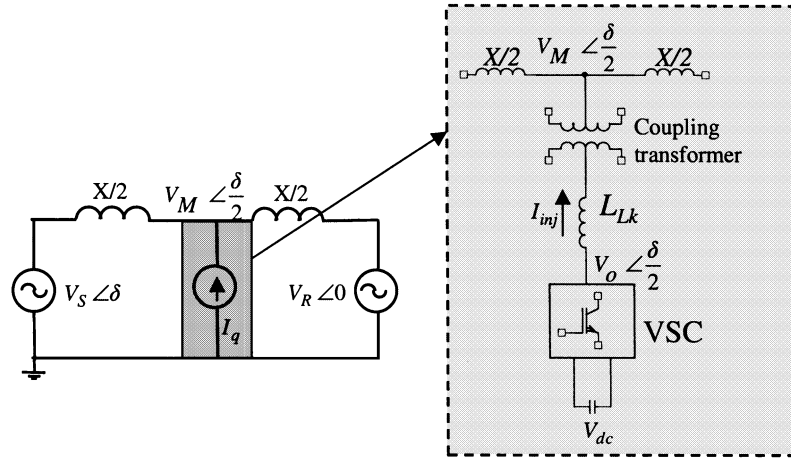
The newer FACTS devices are based on voltage source converters (VSC) implemented using fully controllable devices such as GTO, MCT, IGCT, and IGBT [77]. Within

their voltage and current ratings, the VSC-based FACTS devices are capable of injecting any suitable, controlled voltages and/or currents at the line frequency. The main advantages of these FACTS devices, compared to the thyristor-based devices, are the speed of response and the extended control range, which is mostly independent of the line operating conditions. The main VSC-based FACTS devices are the static compensator (STATCOM), the static synchronous series compensator (SSSC), and the unified power flow controller (UPFC).

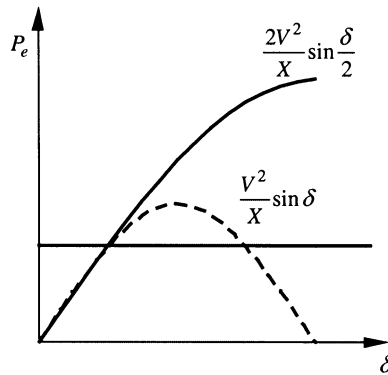
Static Synchronous Compensator (STATCOM). STATCOM is a shunt FACTS device capable of injecting controlled currents at the point of connection with the transmission system [75, 78]. The injected current is usually in phase quadrature (leading or lagging) with the line voltage, so that only reactive power is supplied or consumed by the STATCOM. If real power capability is present, through the use of active energy sources or large energy storage systems, then the injected current can have different phase relationships with the line voltage, thereby extending its control range.

Figure 22-37a shows the schematic of a STATCOM connected at the midpoint of a 2-bus transmission system model. The voltage source converter is capable of generating the required fundamental voltage such that the current injected into the system has the desired phase and magnitude to control power flow. The voltage at the dc link is kept constant by large capacitor banks. Losses in the system are compensated, and the capacitor voltage maintained, by drawing a small real power from the transmission system. The voltage source converter is connected to the transmission line through a line frequency coupling transformer, which enables the STATCOM to work with lower voltage switches.

The output voltage of STATCOM (neglecting losses) is controlled to be in phase with the line voltage. Hence, the system can be modeled as two in-phase, line frequency voltage sources,



(a)



(b)

FIGURE 22-37 STATCOM: (a) midpoint connection, (b) variation of power with phase angle δ .

connected by a reactor (usually the leakage inductance of the coupling transformer), which results in the current I_{inj} being purely reactive. Referring to Fig. 22-37a, if the magnitude of V_o is larger than V_m then the STATCOM feeds reactive power into the system, and if V_o is smaller, it absorbs reactive power.

Referring to Fig. 22-37a, the amplitudes of the sending end, midpoint, and receiving end voltages are assumed to be equal for simplicity ($V_s = V_m = V_r = V$). The STATCOM compensation at the midpoint effectively segments the transmission line into two independent parts, each with an effective line reactance of $X/2$. Neglecting losses, the real power flow is the same in both parts, and can be derived as given in Eq. (22-57).

$$P = \frac{V^2}{(X/2)} \sin \left(\frac{\delta}{2} \right) \quad (22-57)$$

where δ is the angle between the sending and receiving end voltages. Figure 22-37b shows the variation of real power flow with phase angle, as determined by Eq. 22-57. The curve corresponding to

no compensation is also shown for comparison. It can be clearly seen that the STATCOM significantly improves the transient stability margin, that is, for a given fault clearing time, STATCOM allows a much higher real power to be transmitted. In the case of power system oscillations, such as inter area oscillations, the shunt compensation is varied dynamically to provide damping.

Static Synchronous Series Compensator. SSSC is a series connected device that injects a synchronous line frequency voltage, normally in quadrature with the line current. The SSSC controls power flow by controlling the line voltage amplitude, phase angle, and effective line impedance.

Unified Power Flow Controller. The UPFC is a versatile FACTS device that combines the functions of a STATCOM and an SSSC, and extends their capability to inject shunt current or series voltage that involve real power flow as well [79]. With UPFC, the real and reactive power can be controlled independently. Unified power flow controller is capable of controlling all the power system parameters such as voltage magnitudes, phase angles, and effective line impedance simultaneously and therefore meet multiple control objectives. Figure 22-38a shows the schematic diagram of the UPFC. It consists of two voltage source converters with separate controllers but sharing a common dc link with dc storage capacitors.

In present installations of UPFC, most of the control functions are performed by the series converter by injecting a voltage V_{inj} whose phase is independent of the line current and can vary practically from 0° to 360° . The magnitude of the injected voltage can also be varied continuously within the rating of the series converter. The main function of the shunt converter is to provide the real power exchanged by the series converter with the system. It may be noted that the real power exchanged by the series converter is ultimately derived from the transmission line, but the reactive power is

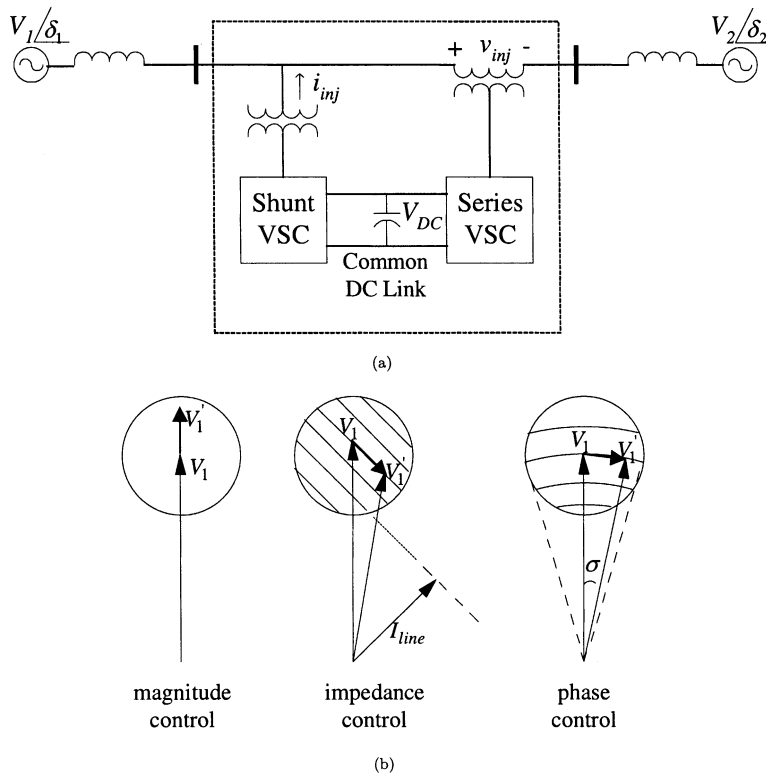


FIGURE 22-38 Unified power flow controller: (a) schematic, (b) control capabilities.

absorbed or supplied locally by the series converter and does not need to come from the transmission system. The shunt converter can be operated at unity power factor or can be controlled to provide additional functions beyond supporting real power needs of the series converter.

Figure 22-38*b* highlights the capabilities of the series converter of the UPFC, namely, control of voltage magnitudes, phase angles, and impedance [75]. With the injected voltage in phase or antiphase with the line voltage the UPFC provides voltage regulation or magnitude control. For a given voltage rating of the series converter, the in-phase addition provides maximum voltage magnitude control. For impedance control, the magnitude of the injected voltage is proportional to the line current and the phase is in quadrature (leading or lagging). In phase angle control, the magnitude and angle of the injected voltage are controlled such that the sending end voltage has the required phase angle without any change in the magnitude. These three features can be combined to achieve multifunction power flow control.

22.10.3 Custom Power

The digital age loads, for example, processing plants in semiconductor industry and data centers, require clean and uninterrupted power. These loads are highly intolerant to (even momentary) power quality problems such as voltage sags or interruptions, harmonics in line voltage, phase unbalance, and flicker in supply voltage. Power electronic systems that mitigate power quality problems in utility distribution systems (1 to 38 kV) are defined as custom power devices [80–82].

Similar to FACTS, the custom power devices can be connected in shunt or series with the distribution line or a combination of both. The major custom power devices are the dynamic voltage restorer (DVR), distribution static compensator (DSTATCOM) and the unified power quality controller (UPQC). The DVR is a series connected device that injects a controlled voltage to compensate for voltage sags and other momentary disturbances. The DSTATCOM is a shunt-connected device injecting controlled currents at the point of common coupling to compensate for power quality problems in the load current. Unified power quality controller combines the features of DVR and DSTATCOM.

Dynamic Voltage Restorer. Short duration voltage sags are the predominant power quality events, with estimated revenue lost per event of more than \$1 million for pharmaceutical industries. Power acceptability curves that quantify voltage disturbances in terms of magnitude of these sags (and swells) and duration of the disturbance have been developed [83]. The most popular of the power acceptability curves is the Computer Business Equipment Manufacturers Association (CBEMA) curve shown in Fig. 22-39. This was developed by the CBEMA, now the Information Technology Industry Council (ITIC). The semiconductor industry has its own standard called SEMI F47, developed by the Semiconductor Equipment and Materials Institute (SEMI).

Dynamic voltage restorers, which are among the most installed custom power devices, protect sensitive equipment against short term voltage disturbances [84]. Figure 22-40 shows the schematic of a DVR. As seen, DVR is a voltage source converter based series connected device that injects a line frequency voltage of appropriate magnitude and phase such that the voltage across the sensitive loads is always well regulated, and any disturbances in the input voltage is not propagated to the load. The voltage source converter is implemented using IGBT switches, which operate at frequencies in the range of tens of kilohertz. They have fairly high control bandwidth and can respond to voltage disturbances in a small fraction of the line frequency cycle.

When the DVR is not connected to an active dc source and cannot handle real power in steady state, the injected voltage is constrained to be in phase quadrature with the load current. With this mode of control, the magnitude of sags that a DVR can correct becomes a function of the load power factor, and at higher power factor (close to unity) only a smaller voltage disturbance can be corrected. Since most of the sag events are of short duration, many of the installed DVRs rely on the large energy storage capacitor to supply real power for a short duration, and not constrain the injected voltage to be in phase quadrature. Several other installations use a separate rectifier to supply real power to the dc link from the distribution system [85].

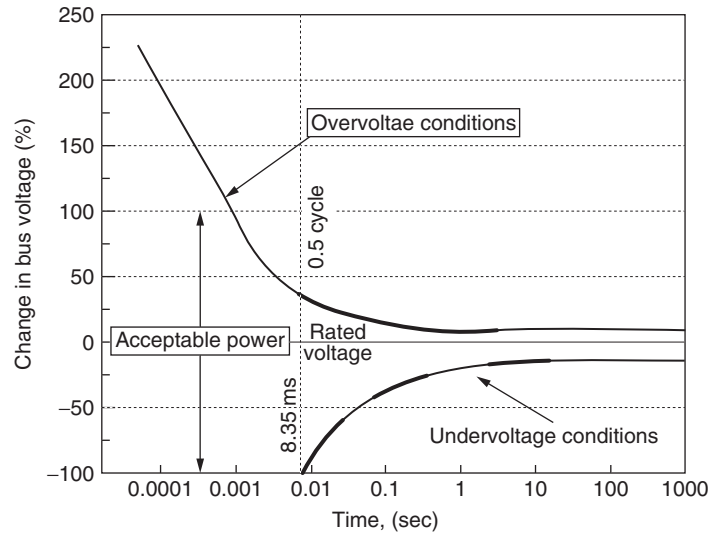


FIGURE 22-39 The CBEMA curve.

Distribution Static Compensator. The Distribution STATCOM has similar structure as that of the STATCOM used in transmission systems, and injects controlled currents. However, the main objectives of DSTATCOM are quite different. The load currents in distribution system can be unbalanced and contain reactive and harmonic components. Standards such as IEEE 519 and IEC 61000 place limits on maximum permissible harmonic currents for various types of equipment and voltage levels [47, 48]. The DSTATCOM with closed loop control injects correction currents such that the compensated load draws balanced, fundamental, unity power factor current.

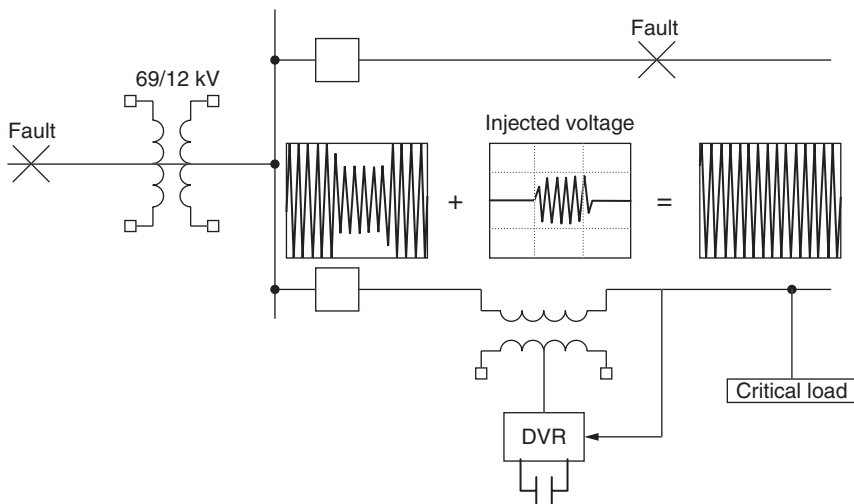


FIGURE 22-40 Application of a DVR.

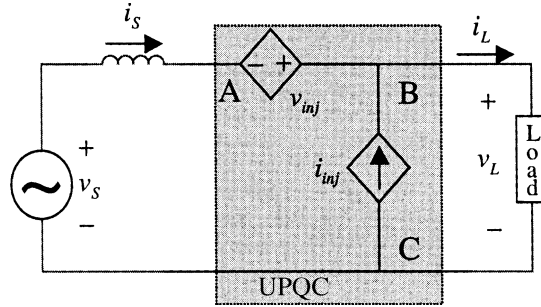


FIGURE 22-41 Unified power quality conditioner.

Unified Power Quality Conditioner. The UPQC combines the features of a DVR and DSTATCOM and can inject current in shunt and voltage in series simultaneously. Figure 22-41 shows the schematic of a UPQC. It has the same structure as that of UPFC used in transmission systems, consisting of two voltage source converters sharing a common dc link. One of the converters is connected in series with the distribution line injecting controlled voltages and the other converter is connected in shunt and injects controlled currents. Therefore, the UPQC can simultaneously correct for unbalances and distortion in line voltage as well as load currents.

Solid-state switches used to connect critical loads to multiple feeders or to break short circuit currents, hence improving power quality, are also considered as custom power devices. These are referred to as network reconfiguring devices and include solid-state current limiter (SSCL), solid-state breaker (SSB) and solid-state transfer switch (SSTS) [80]. These are much faster than the conventional mechanical switches and hence significantly enhance the reliability of the distribution system.

22.10.4 Distribution Generation Interface

Distributed generation (DG) is an important energy option for the twenty-first century and a key element of the restructuring of electric grid. Distribution generation, using renewable sources such as solar, wind, and tidal energy, and hydrogen (with photovoltaics for hydrogen generation) represents one of the most promising paths to sustainable energy.

Fuel cells, photovoltaics (PV), wind energy, and microturbines are among the most promising distributed generation technologies at present. Most of the distributed energy resources (DER) require a power electronic converter to interface with the power system network. Fuel cells and PV require dc-ac conversion, microturbines require high frequency to line frequency conversion, and generators used to capture wind energy are controlled through a rotor side power converter. Further, these DER may also require power electronics controlled energy storage.

Figure 22-42 shows a schematic diagram of the converter that can be used to interconnect photovoltaics with the utility. The input voltage from the solar cell array, typically 52 to 90 V, is converted to a higher magnitude, well regulated and isolated dc voltage through a high-frequency dc-dc converter. The dc link voltage is then converted to the required 60 Hz ac voltage by using a PWM

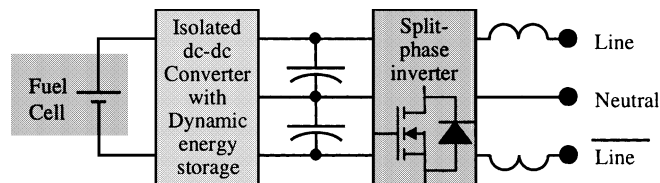


FIGURE 22-42 Fuel cell connection to the grid.

voltage source converter. An important feature of a PV interface is the maximum power point tracking circuitry designed to derive the maximum possible energy from solar radiation by suitably adjusting the current drawn from the solar panel. Fuel cells can also be interfaced to the grid by using a similar configuration as shown in Fig. 22-42. Since fuel cells have very little overcurrent or short-circuit rating, an energy storage capacitor is installed at the input side of the converter.

Wind energy is widely considered to be the fastest growing alternate energy source. The cost of wind energy in large wind farms located in good wind sites, at about 5 cents/kWh, is now competitive with the conventional utility generation, and is expected to reduce further. In 2004, the total installed capacity of wind energy worldwide was above 46,000 MW, with the United States accounting for 7000 MW [86].

Large wind farms consist of several multi megawatt wind turbines that are interconnected with the utility grid through medium voltage collector network. Doubly fed induction generators (DFIG) with a wound rotor and an ac-dc-ac PWM converter, as shown in Fig. 22-43, is becoming the most widely used technology for wind generation. The main advantage of the DFIG based generation is that it allows extraction of maximum energy from the wind at varying wind speeds, with the power converter rated only for about 15% of the total power. The stator winding is connected directly to the utility grid while the rotor is supplied with controlled, variable frequency currents by the PWM converter. By appropriate control of the rotor currents, the machine can generate power from sub-synchronous to super synchronous speeds. Another advantage of DFIG is that the grid side converter can generate or absorb reactive power.

Apart from serving as an environment-friendly energy source, the DG systems are expected to provide various other benefits. For example, the concept of microgrids is gaining prominence [87]. Microgrid is a cluster of DER with power converters, energy storage and loads, which can be controlled together and present to the grid as a single entity. With suitably designed power converters, and with coordinated control, the microgrids can enhance stability, provide relief to transmission congestion, and provide reactive power support. Another promising approach is to use distributed micro sources for combined heat and power (CHP) [88]. Due to the various ancillary functions expected, inverter based DG are becoming more widespread compared to traditional reciprocating engines.

The interconnect standards for DG are just being evolved, like the IEEE 1547 standard for interconnecting distributed resources with electric power systems [89]. IEEE 1547 specifies the requirements for the DG to disconnect from the grid under deviations in voltage magnitude and frequency or under grid outages. Much work is still needed to realize a universal interconnect technology (UIT) and to understand the effect of large penetration of DG on fault currents, protection and dynamic interactions with the power system.

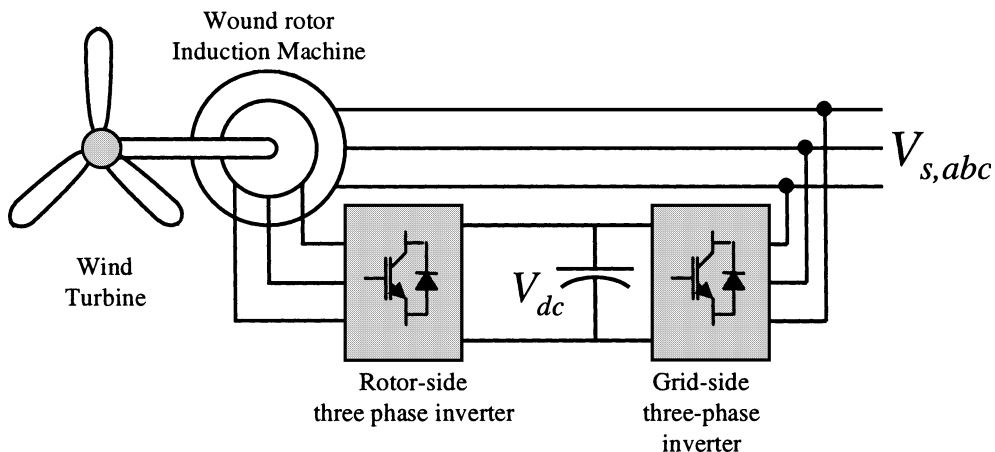


FIGURE 22-43 Grid connection of a doubly fed induction generator.

22.11 COMPONENTS OF POWER ELECTRONIC CONVERTERS

This section describes the individual components that constitute a power converter. These include power semiconductor devices and passive components (inductors, transformers, capacitors). A detailed description of the structure and physics of power semiconductor devices is beyond the scope of this discussion. The interested reader is referred to standard books on this subject [4, 90, 91]. Details of magnetics, material properties, and capacitors are also omitted. Unlike semiconductor devices and capacitors, very few magnetics are available as standard products from manufacturers, and therefore require custom design. Magnetics design is covered in significant detail in Ref. [92]. For details of capacitors, application notes and datasheets supplied by manufacturers have to be relied on.

22.11.1 Power Semiconductor Devices

Power electronic circuits require high-power semiconductor switches and diodes. An ideal switch should have the following characteristics: full control over switch state (on/off), very low voltage drop during on state, infinite impedance during off-state, and instantaneous transition between states. Diodes should have very low voltage drop during conduction, infinite impedance in off-state, and instantaneous transition. Practical devices have nonideal characteristics, and different devices capitalizing on one advantage while sacrificing some other have been developed. Figure 22-44 shows the circuit symbols of common power semiconductor devices. These are listed below with their voltage, current, and switching limitations.

Diodes: Line frequency, fast recovery, ultra-fast recovery, and schottky—in increasing order of switching speed, and decreasing order of reverse voltage rating.

MOSFETs (metal oxide semiconductor field effect transistor): Good for low voltage ~100s of volts, high switching frequency (> 100 kHz).

IGBTs (insulated gate bipolar transistors): Good from a few hundred Volts to about 6 kV, currents upto 1.2 kA, and switching frequency upto 30 kHz.

Thyristors or SCRs (silicon-controlled rectifiers): Good for very high voltage and current (~kV and kA), and low power moderate performance applications.

GTOs (gate turn-off thyristors): Good for very high voltage (~kV), high current applications (~kA), with switching frequency upto a few kilohertz.

Miscellaneous: IGCT (integrated gate commutated thyristor), MCT (MOS-controlled thyristor), BJTs (bipolar junction transistors), triacs, diacs.

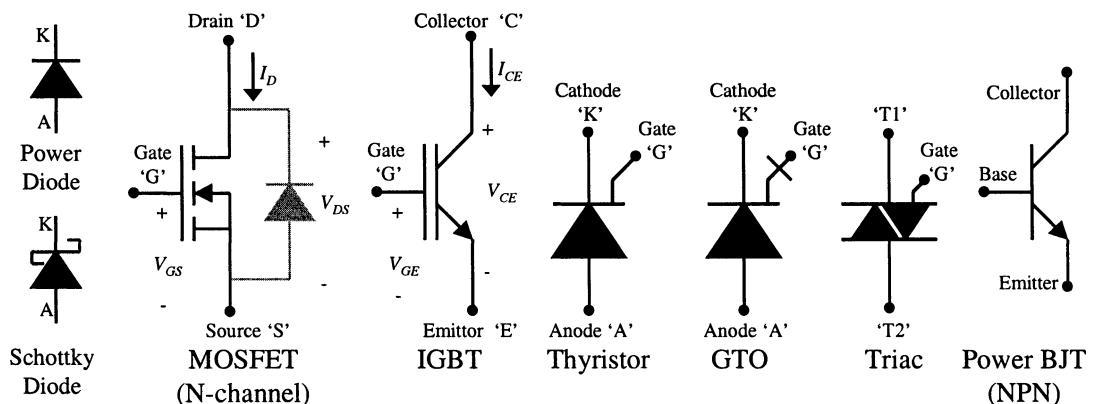


FIGURE 22-44 Circuit symbols of common power semiconductor devices.

Diodes. There are essentially two types of power semiconductor diodes: PN junction, and metal-semiconductor junction (schottky). PN junction power diodes have an additional N^- (lightly doped with N type impurities) drift region, so the overall structure is PN^-N . The depletion layer of the PN junction extends in the N^- region when the diode is reverse biased, and its length determines the maximum reverse voltage the diode can block. Intrinsically, the N^- region has a high resistance. However, when the diode is forward biased there is injection of excess carriers in this region resulting in a low effective resistance. This phenomena is commonly called *conductivity modulation*. The on-state voltage drop across the diode consists of the PN junction drop and the resistive drop in the drift region; typically it ranges from 0.6 to 1 V. There is a small delay in going from off-state to on-state due to the time required for the carriers to build up. During turn-off, the excess carriers in the drift region have to be removed. Thus, for a short time, the diode conducts in the reverse direction with a high voltage across it. This phenomena, known as reverse recovery, leads to significant power loss and becomes one of the limiting factors in high-frequency circuits. PN junction type power diodes are classified as:

- line frequency rectifiers: for rectification of 50/60 Hz utility input.
- fast and ultra-fast recovery diodes: for high frequency rectification. These have recovery time ranging from a couple of μs to fractions of μs .

Schottky diodes are based on metal-semiconductor junctions. These junctions have a lower junction potential leading to a lower forward voltage drop. Silicon-based schottky power diodes have forward voltage drop ranging from 0.3 to 0.6 V, and can withstand reverse voltages up to 200 V. As opposed to PN junction diodes, schottkys are majority carrier devices, so they do not have any reverse recovery. However, compared to PN junction power diodes, they have significantly higher capacitance. This capacitance, in combination with circuit inductances, can lead to significant oscillations when the diode goes from on-state to off-state. Silicon-based schottky diodes are suited for very high frequency, low voltage, and high current rectification. Recently, SiC-based schottky diodes have been developed and are now commercially available with rating up to 1200 V [93, 94]. Even though SiC diodes have a much higher forward voltage drop (2 to 3 V), the absence of reverse recovery makes them suitable for high voltage high frequency rectification.

MOSFETs. Unlike signal level MOS devices that are fabricated laterally, power MOSFETs have a vertically diffused structure. For an N-channel MOSFET the doping is of the form $N^+PN^-N^+$. The drain is the N^+ terminal next to the N^- region, while the source is the N^+ region next to the P region. A positive voltage on the isolated gate terminal produces an electron channel in the P-region allowing current to flow from drain to source (or vice versa). Further, the P-type body is shorted to the source terminal resulting in an intrinsic diode with anode at the source and cathode at the drain. Although this diode is not very good in performance, it is useful for most power electronic circuits. Both P and N channel power MOSFETs are available, but N-channel MOSFETs are more prevalent due to their lower on-state resistance. Our discussion will therefore be restricted to N-channel MOSFETs.

The steady-state V-I characteristics for an N-channel MOSFET are shown in Fig. 22-45a. For $V_{GS} < V_{th}$, the MOSFET acts as an open circuit from drain to source; V_{th} , the threshold voltage, is in the range of 2 to 4 V. For $V_{GS} > V_{th}$, the MOSFET follows the characteristics shown in Fig. 22-45a. In amplifier circuits, MOSFETs are operated in their active region, where the drain current I_D is almost independent of the drain to source voltage V_{DS} . Power MOSFETs are operated in the ohmic region where I_D is proportional to V_{DS} , and the MOSFET behaves like a resistance. The effective on-resistance, designated R_{DS} , depends on V_{GS} , I_D , and the junction temperature T_j . For MOSFETs, R_{DS} increases with increasing temperature, a property useful in paralleling of devices to obtain higher current carrying capacity. The maximum value of V_{GS} is usually ± 20 V; for logic level power MOSFETs, it is limited to ± 10 V. For most MOSFETs, increasing V_{GS} beyond 10 V does not have significant effect on R_{DS} . MOSFETs are rated for (1) maximum drain to source breakdown voltage (BV_{DSS}) (2) maximum continuous average current for a specified temperature (e.g., I_{D25} at 25°C), and (3) a safe operating area (SOA) in terms of V_{DS} , I_D , and time duration for which I_D flows. In very high current applications, thermal performance of the device package can also impose an additional constraint.

MOSFET gate drive circuits usually run on a nominal $V_{cc} = 15$ or 12 V, and switch V_{GS} between V_{cc} and 0 V. For very high current applications and improved noise sensitivity, a negative V_{GS} may

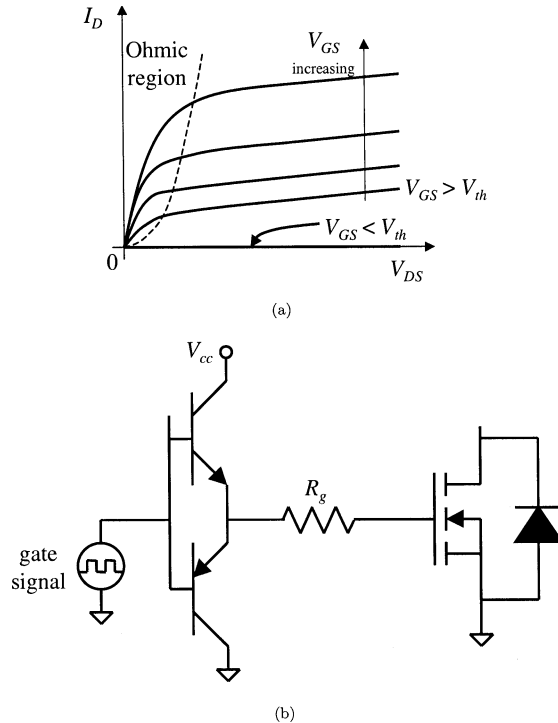


FIGURE 22-45 MOSFETs: (a) V-I characteristics, and (b) simple gate drive circuit.

be applied during the off-time. Since the gate is insulated from the source there is no dc current flow from gate to source. However, depending on their ratings power MOSFETs have a significant input capacitance C_{iss} . Thus, to increase V_{GS} from 0 to 15 V in a very short time (20 to 100 ns) a significant current (order of 1 A) is required. Figure 22-45b shows a basic MOSFET gate drive circuit. Several gate drive ICs that can source and sink current up to a few amperes, while providing several other auxiliary functions, are commercially available [9, 12]. For *high-side switches*, switches whose source voltage changes with the switch state, as is the case for the MOSFET in a buck converter, an isolated gate drive is required. The isolation is provided by using either high-frequency transformers or high-speed optocouplers. There are several standard gate drive configurations, each with its own pros and cons. The best source of these circuits are application notes from device manufacturers (e.g., see Ref. [62]).

Power MOSFETs can be used up to a few 100 kHz and in some applications in the megahertz range. To increase operating frequencies further, *RF Power MOSFETs* have recently been introduced [95]. These can be operated in the 10 MHz range and are expected to reduce converter size and weight considerably.

Insulated Gate Bipolar Transistors. The structure of IGBTs is similar to MOSFETs. However, IGBTs have significantly higher voltage and current ratings compared to MOSFETs, and their on-state voltage drop is also lower. In addition, several IGBT chips are paralleled inside one package to form an IGBT module with a significantly higher current rating. Insulated gate bipolar transistors modules are commercially available in voltage rating up to 6.5 kV, and currents up to 1200 A. The gate drive requirements of an IGBT are similar to that of MOSFETs. It is turned on by applying a positive gate source voltage (typically 15 V), and turned-off by applying a smaller negative voltage

(about -5 V). Compared to MOSFETs, IGBTs have much longer switching times. Insulated gate bipolar transistors are minority carrier devices, so their turn-off is characterized by a “tail current”. Their switching frequencies are generally limited to 30 kHz, and the maximum switching frequency reduces with power level. If resonant power conversion techniques are used, the tail current during turn-off can be avoided and the switching frequency can be pushed higher.

Thyristor and Similar Devices. Thyristors, also called silicon-controlled rectifiers (SCRs), are high-power semiconductor devices that can block voltage of either polarity and conduct current in one direction only (from anode to cathode). They can be switched on by applying a current pulse to their gate terminal (with return path through cathode) when there is a positive voltage from anode to cathode. They can be switched off only by reducing the device current to zero. Thyristors are available in very high current and voltage ratings and have a very low conduction voltage drop. Thyristors are mostly used for ac to dc (or vice-versa) power conversion, where the device current is expected to reduce to zero. Normally, thyristors are switched at or close to the ac system frequency. They are ideally suited for utility applications such as HVDC and controlled reactors.

Gate turn-off thyristors are very high power devices, with their low-end ratings overlapping with the high-end rating of IGBTs. Unlike a thyristor, they can be turned on and off using the gate terminal, although the gate drive is more complex compared to a MOSFET or an IGBT. The switching times for GTOs are of the order of $10\ \mu\text{s}$ so their maximum switching frequency is in the kilohertz range. They are used exclusively in very high power applications like some motor drives, FACTS devices, and active filters. Gate turn-off are commercially available in ratings upto 6 kV and 6 kA. Enhancements to GTO have led to the development of IGCT (integrated gate commutated thyristor) [96] or GCT (gate commutated turn-off thyristor) [97], ETO (emitter turn-off thyristor), and MTO (MOS turn-off thyristor) [91]. At present there is a strong research effort in development of high power devices with high switching speeds. The impetus behind these are high power pulse applications for defence, and the increasing utility applications of power converters.

A “back-to-back connection” of two thyristors (anode of one to the cathode of other) can conduct current in either direction and block voltages of either polarity. Triacs realize the functionality in a single semiconductor device. However, voltage and current ratings of triacs are very low compared to that of thyristors.

Power *bipolar junction transistors* (BJTs) have been almost completely replaced by MOSFETs and IGBTs, due to ease of control and higher switching frequency. However, they are still used in some applications like linear power supplies.

22.11.2 Magnetic Components

In power electronic converters three types of magnetic components are used: single winding inductors (for filtering current and aiding in resonant transitions in some circuits), multi-winding coupled inductors (to provide filtering and isolation), and transformers (for isolation and stepping up/down voltage). Unlike semiconductor devices, these have to be custom designed using available cores, wires, etc. The primary consideration for their design are size/weight and power loss. Design procedures using the common *area product method* is presented here. Another approach is the *core geometry method*, which can be found in Refs. [5, 92]. It is assumed that the reader is familiar with basics of electromagnetism and magnetic circuits. An E-E type core will be used for illustration, but the method is applicable to any core shape. The E-E type core construction, along with relevant definitions of its geometry, is shown in Fig. 22-46. The complete core is formed with two E cores with possibly an air gap between them. The coils are wound over a plastic bobbin placed on the outside of the center leg of the E sections.

Due to high-frequency magnetic fields, significant eddy currents are induced in the magnetic core and windings, and due to high-frequency electric field there can be significant capacitive currents between windings. Eddy currents lead to significant losses and necessitate the use of high-resistivity magnetic core materials and thinner wires (which may be paralleled for high current capacity) for the coils. Loss due to magnetic hysteresis also increases with increase in frequency.

Magnetic Core Materials. Inductors use one of three different kinds of core materials, depending on the currents they are supposed to carry. The three core material types are silicon steel for low-frequency filtering, powdered iron for high-frequency filtering, and ferrites for carrying high-frequency currents. Transformers in power converters always carry high-frequency currents and therefore use ferrite cores.

For low-frequency filtering, where the inductor primarily carries a dc component with a 120 Hz ripple (as in line frequency rectifiers), standard silicon steel laminations can be used to form the core. Silicon steel has high saturation flux density (~ 1.8 Tesla) and high relative permeability ($\mu_r \sim 40,000$). This results in a small core size without incurring significant core loss. However, the relative permeability reduces with increase in flux density, so the effective inductance value changes with the dc component of the current.

For inductors that are supposed to carry currents consisting of a large dc component and a small high-frequency ac component (e.g., the inductor used at the output of a dc-dc converter), powdered iron or MolyPermalloy Powder (MPP) cores are used. These cores have high resistivity leading to lower eddy current losses, high saturation flux density (upto 1.4 Tesla) leading to lower core size, and distributed air gap leading to a low relative permeability. As with silicon steel, the relative permeability reduces with increasing flux density. The cores are available in toroidal and E shapes. Toroids are difficult to wind but can be used to make very good quality inductors.

For transformers and inductors that have to carry significant high-frequency current, the powdered iron material has unacceptable loss due to eddy currents and hysteresis. Ferrite materials, with very high resistivity, lower weight, but low saturation flux density (~ 0.4 Tesla) are suitable for these. Common ferrite materials are 3F3 from ferroxcube and PC44 from TDK. Cores made from these materials are available in a wide variety of shapes and sizes: toroids, pot core, E, PQ, RM, etc.

Inductor Design. For inductors, the design requires specified values of the inductance (L), and the peak and rms current ($\hat{I}$ and I_{rms}) it has to carry. The relations between peak current and core area (A_{core}), rms current and window area (A_w), and the expression for the *area product*, $A_p = A_{core} \cdot A_w$, are as follows:

$$\lambda = L \hat{I} = N \Phi = N B_{max} \cdot A_{core} \quad (22-58)$$

$$\Rightarrow A_{core} = \frac{L \hat{I}}{N B_{max}}$$

$$I_{rms} = J \cdot A_{cond} = J \cdot \frac{A_w \cdot k_w}{N} \quad (22-59)$$

$$\Rightarrow A_w = \frac{N I_{rms}}{J k_w}$$

$$A_p = A_w \cdot A_{core} = \frac{L \hat{I}_{rms}}{J B_{max} k_w} \quad (22-60)$$

Here, λ is the flux linkage, N is the number of turns of the winding, Φ is the magnetic flux corresponding to peak current $\hat{I}$, B_{max} is the maximum flux density (may be different from the saturation

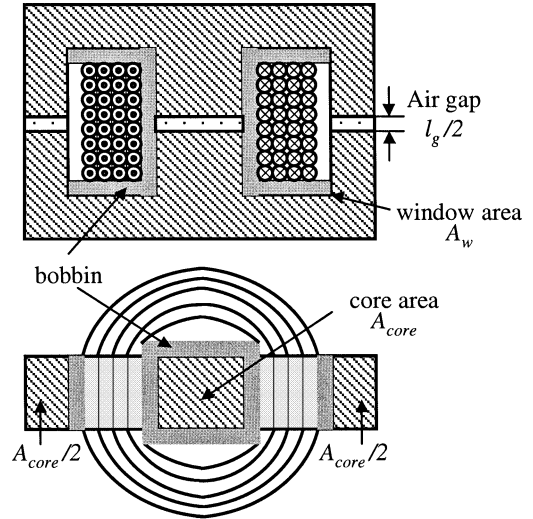


FIGURE 22-46 Vertical and horizontal sections of an E-E core inductor.

flux density), J is the current density in the winding, A_{cond} is the cross-sectional area of one conductor in the winding, and k_w , called the winding factor is an empirical factor used to indicate the fraction of window area utilized by the copper of the windings. In the expression for A_p , the numerator consists of specified quantities, while the denominator has quantities which are chosen for design. B_{max} may be chosen close to the saturation flux density depending on allowable core loss; J is chosen in the range of 2.5 to 8 A/mm², depending on allowable conduction loss, mechanism for heat removal, and the allowable temperature rise; k_w is in the range of 0.3 to 0.8 depending on space taken by bobbin and insulation, and the method of winding. Once the area product has been calculated, a core with the desired value of A_p can be chosen from manufacturers' catalogs. The number of turns and the air gap required to achieve the desired inductance value are then calculated as

$$N = \frac{L \hat{I}}{B_{max} A_{core}} \quad (22-61)$$

$$L = N^2 / \mathfrak{R} = N^2 \mu_o A_{core} / l_g \quad (22-62)$$

$$\Rightarrow l_g = N^2 \mu_o A_{core} / L \quad (22-63)$$

where $\mathfrak{R}$ is the core reluctance. The above equation assumes that $\mu_r / l_m \gg 1 / l_g$, where l_m is the magnetic path length of the core. The conductor size is calculated as $A_{cond} = k_w \cdot A_w / N$. If the current is expected to have significant high frequency component, the radius of the conductors should not exceed the *skin depth* of copper at the expected frequency. Several conductors may then be paralleled to achieve the required value of A_{cond} . Finally, P_{cond} , conduction loss in the windings, and P_{core} , loss in the core, should be calculated. P_{cond} requires an estimate of the winding resistance, while P_{core} is computed using manufacturer supplied empirical loss curves. Core loss in an inductor depends on the operating flux density B (related to the average value of the current), and change in flux density during each switching cycle ΔB (related to the expected current ripple). It is generally accepted that $P_{cond} = P_{core}$ indicates a good design. However, power loss in the core can be dissipated more easily compared to windings (especially if there are several layers). Thus, it may be desirable to have $P_{cond} < P_{core}$. This affects the choice of B_{max} and J values used above.

Powdered iron cores have distributed air gaps and the above procedure is not directly applicable to them. For selection of these cores, the manufacturers suggest a number of cores based on a metric similar to the product $L \hat{I}_{rms}$. The final core selection can then be made based on allowable losses (P_{cond} relating to J , and P_{core} relating to B_{max}), and temperature rise. The cores have a specified A_L value, the inductance obtained from the core in nH/Turns². The number of turns required to obtain the inductance value can be computed using the specified A_L value. Finally, the inductance variation with dc current should be checked to make sure that the required value is obtained at the maximum operating current.

Transformer Design. For a transformer, the design specifications are the turns ratio, rms current in all the windings, and the maximum volt-second product ($[V \text{ sec}]$, the maximum product of voltage and time for which the voltage is applied) expected across the primary winding. The window area can be related to the rms current, the core area to the maximum volt-second product, and the resulting area product calculated as follows

$$k_w \cdot A_w = N_p \cdot A_{cond,p} + \sum_m (N_{sm} \cdot A_{cond,s}) \quad (22-64)$$

$$\Rightarrow A_w = \frac{N_p}{J k_w} \cdot \left(I_{p,rms} + \sum_m [(N_{sm} / N_p) \cdot I_{sm,rms}] \right)$$

$$\Delta \lambda = [V \text{ sec}] = N_p \Delta \Phi = N_p \Delta B_{max} \cdot A_{core}$$

$$\Rightarrow A_{core} = \frac{[V \text{ sec}]}{N_p \Delta B_{max}} \quad (22-65)$$

$$A_p = A_w \cdot A_{core} = \frac{[V \text{ sec}] \cdot \left(I_{p,rms} + \sum_m [(N_{sm} / N_p) \cdot I_{sm,rms}] \right)}{J \Delta B_{max} k_w} \quad (22-66)$$

The subscripts “p” and “s” indicate primary and secondary, respectively, while the index m refers to the secondary number. In the above equations, ΔB_{max} is the maximum *change* in flux density; this equals the maximum value of B_{max} for single quadrant operation (as in a forward converter), and twice B_{max} when flux direction reverses (as in full-bridge dc-dc converters). Core loss has more than quadratic dependence on ΔB_{max} . For transformers, the value of B_{max} is therefore chosen much lower than the saturation flux density (about 0.1 to 0.25 T). After a core is chosen by utilizing the computed A_p , the number of primary turns are calculated as

$$N_p = \frac{[V \text{ sec}]}{B_{max} A_{core}} \quad (22-67)$$

In high-frequency transformers, the magnetizing current can be a significant part of the primary current and should be accounted for in the above calculations. Conductor selection for transformers is similar to that for inductors. For very high frequency and high current applications use of insulated copper foils and *Litz wire* are alternatives to use of multiple parallel conductors.

It should be noted that the procedures described above are rudimentary and several other factors have to be accounted for in a real design. These include ambient temperature, temperature rise of the core (which affects the saturation flux density and the core loss significantly), *hot-spot* temperature, winding loss due to *fringing fields* near the air gap, and winding loss due to *proximity effect*. In transformers, interleaving primary and secondary windings helps in reducing the proximity loss and leakage inductance at the expense of increased interwinding capacitance. To eliminate the winding process, planar magnetics that use tracks on multilayer PCBs to form the windings have been developed. Planar magnetics also have the benefit of low profile. To reduce the number of discrete magnetic components, there has been significant research in integrated magnetics—realizing the functionality of several magnetic components using a single magnetic component.

22.11.3 Capacitors

There are several types of capacitors each with different advantages and limitations. For power electronic circuits the choice criteria are: capacitance value, voltage rating, effective series resistance (ESR) and effective series inductance (ESL), and the ripple and ac current capacity at different frequencies. All these can be found as specifications in manufacturers datasheets. ESR is either specified as a number at different frequencies (typically 120 Hz and 20/100 kHz), or in terms of the *dissipation factor*, $\tan\delta = ESR/(2 \times f \times C)$. If ESL is specified, it may be either as a number or in terms of the series resonance frequency, $f_r = 1/(2\pi\sqrt{ESL \times C})$. The main capacitor types used in power electronic circuits are listed in the Table 22-3. The list is by no means exhaustive and there are several other capacitor types suited to different applications, for example, tantalum, mica, and high power film capacitors.

22.11.4 Snubber Circuits

As described in Sec. 22.2.4, there is significant energy loss in power semiconductor devices during turn-on and turn-off. In addition to the phenomena described there, reverse recovery in PN junction based diodes and capacitance of schottky diodes, output capacitance of MOSFETs and IGBTs, and leakage inductance of transformers also contribute to power loss during switching. Snubber circuits are used to remove the transient high voltage and high current stress from semiconductor devices. They can be classified into two basic types—*turn-on* and *turn-off* snubbers, depending on the transition where they are active. With a turn-off snubber, an alternate path is provided for the current during turn-off so that the switch ceases to carry current as soon as it is disabled. This does not necessarily imply that the switching loss is eliminated. In low-power applications, the energy may be dissipated in an external resistor and in fact the energy dissipated may be higher. For high-power applications, the energy may be recovered using energy recovery snubbers, resulting in significantly improved

TABLE 22-3 Capacitors Types Commonly Used in Power Electronic Circuits

Capacitor Type	Characteristics	Application
Electrolytic	Polarized (unipolar voltage only) High density, High ESR and ESL Low reliability Voltage ratings up to 500 V Capacitance up to 100s of mF Rated for ripple current capacity	DC input and output filters, very short time energy storage
Ceramic and Multilayer Ceramic	Very low ESL and ESR Low capacitance values (up to $\sim 100 \mu\text{F}$) Maximum capacitance value reduces with increasing voltage rating	Paralleled for low voltage output filtering, “bypass” for gate drives, high voltage capacitors (up to 1 kV) used in snubber circuits
(Metallized) Polyester film	Low ESR and ESL Low capacitance values (up to a few μF) High voltage ratings (few kV) Bigger than ceramic and electrolytic	In input and output dc filters: to suppress switching transients, ac filters, snubber circuits
Metallized Polypropylene film	Very low ESR and ESL Low capacitance values (up to a few μF) Very high rms current and high voltage ratings	In resonant converters: for carrying high frequency ac current

efficiency at the expense of increased component count. Another use of snubber circuits is to limit transient overvoltages caused due to parasitics.

22.11.5 Heat Sinks

The main power loss in power electronic circuits occurs in power semiconductor devices, windings, and cores of the magnetic components, capacitors subjected to high-ripple current, and auxiliary circuits like gate drives. The resulting heat has to be removed from the components and eventually transferred to the atmosphere/ambient surroundings due to the following reasons. Power semiconductors have a maximum junction temperature rating beyond which they fail, and on-state resistance of devices like MOSFETs increases with junction temperature leading to reduced efficiency. Electrolytic capacitors can fail when heated beyond their ratings and their expected lifetime reduces with increase in temperature. Increased temperatures in magnetics can lead to higher power loss, poor magnetic core characteristics, insulation breakdown, and shorter lifetimes.

Most semiconductor devices are cooled by mounting their package on a heat sink. A heat sink is essentially a piece of metal designed to dissipate heat by convection and radiation (by maximizing the heat sink surface area). The heat transfer from the device to the heat sink is via conduction. If the package of the device is not electrically isolated from its electrical terminals, an interface material (like Silpads) with good thermal conductivity and high dielectric breakdown voltage has to be used. For magnetics, resins may be used to conduct heat from the windings and the core to the metal heat sink. The heat sink itself may be cooled naturally, by forced convection (using fans to blow air on the heat sink surface), or liquid cooled (by circulating a liquid through the mass of the heat sink, and cooling the liquid by a radiator). Simple steady analysis and design of heat sinks is carried out using specified *thermal resistances*, ratio of temperature difference to heat transferred across the material, of all the materials in the heat transfer path. For pulsed applications, a transient thermal model, usually consisting of first-order lags between two consecutive interfaces, is used.

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